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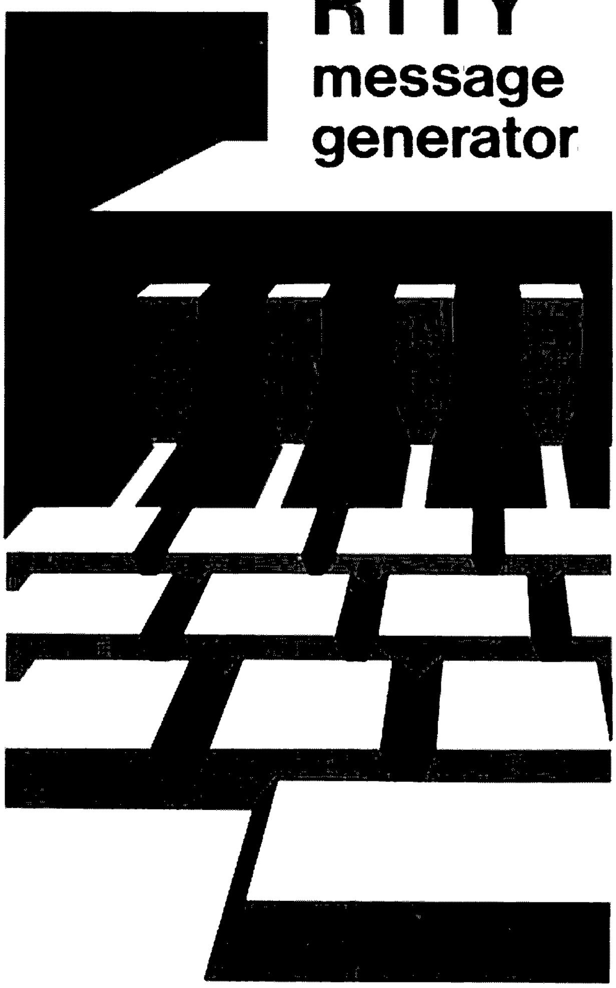
random access memory RTTY message generator



this month

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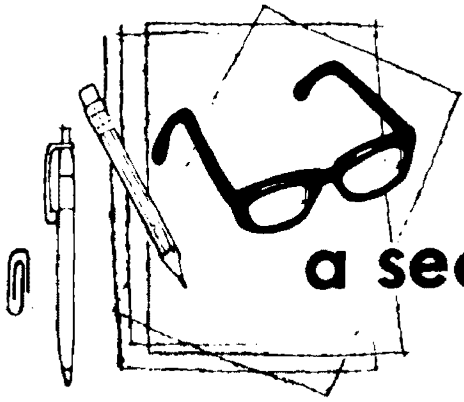


a second look

by jim fisk

With more than 50-million hand-held digital calculators already in the hands of consumers, and predicted sales of another 34-million units in 1975, it's easy to understand why the simple four-function machines that sold for nearly $100 only two or three years ago are now going for $25 or so – less at large discount outlets. The prices have been slashed on the more sophisticated scientific calculators, too, and with each new model that is introduced the price of the models it replaces falls still further. The Hewlett-Packard HP-35 Electronic Slide Rule, for example, originally sold for $395, but when the more sophisticated HP-45 appeared the price was cut to $295. It has since dropped to $225 with some discount stores selling it for even less. Similar price cuts (or larger) have been posted by practically all the manufacturers.

New integrated circuits which were on the drawing boards at major semiconductor manufacturers in October and November, 1974, promise a whole new generation of scientific calculators will be available within the next few months. Price and performance breakthroughs in this fast-moving industry have generally lagged the development of new calculator chips by only three or four months, and since several new circuits are scheduled to appear this month or next, you can expect to see newer, smarter electronic slide rules on the market by early summer. Many of their features are still closely guarded secrets, but look for basic scientific calculators capable of evaluating expressions that include transcendental functions and one or more levels of parenthesis for less – perhaps substantially less – than $100.

Typical of the new calculator circuits is the two- or three-chip Senior Scientist set introduced by MOS Technology in late 1974 which makes it possible to build a more powerful non-programmable hand-held calculator than any now available. Recently this firm also introduced single-chip, 28-pin scientific calculator arrays which offer full scientific functions, algebraic entry, two parenthesis levels and full feature memory. Another of the new calculator chips is the Rockwell Microelectronics 4802, which will go into full production this month and will be priced at under $20. It offers, among other things, degree-radian conversion and factorials.

Programmability, which is now available only in relatively expensive calculators such as the Hewlett-Packard HP-65 ($795) and the Monroe/Compucorp Beta 326 ($695) should be available in other pocket-sized calculators in the $200 to $300 price range sometime later this year. However, don't expect these machines to have the sophisticated magnetic card reader featured in the HP-65 – they will probably require that you key in the program each time you wish to use it. Future models will probably overcome that small detail, and will probably offer interface with a variety of peripheral devices such as printers as well, but with the powerful, multi-function calculators that are available now at reasonable prices, can you afford to wait?

Jim Fisk, W1DTY
editor-in-chief

presstop

OSCAR 7 LAUNCH WAS PERFECT, and it ended up where it was supposed to be when it was supposed to be there. Both telemetry beacons are working perfectly, and the only hitch is a marked lack of sensitivity in the 2- to 10-meter translator. The 432-MHz to 2-meter translator, on the other hand, is working much better than expected -- predictions were that 350 W effective radiated power would be required for a consistent signal through it, and it's doing so well that a maximum of 80 W ERP is being requested!

A Number of European Stations have been copied in the midwest coming through the translator from 432, and despite a somewhat narrower than expected bandwidth (about 80 kHz) the efficiency is quite high. The two-meter beacon is somewhat lower in frequency than expected -- despite uncertainties from doppler shift, it appears to be about 145.972 instead of 145.980 MHz.

The Two-To-Ten Meter Translator is a disappointment, so far. It takes a lot more signal to get into than Oscar 6, and very deep fades are frequent during its passes. The beacon is very strong, and bandwidth is considerably greater than the planned 100 kHz (signals as high as 29520 kHz have been copied thru it).

Complete Computed Orbital Data for Oscars 6 and 7 will be available monthly as an added slip-in sheet to HR Report. Copies of these predictions will be provided to interested readers upon receipt of an SASE (one SASE for each month).

AMATEURS RESPONSIBLE FOR "CODED" TRANSMISSIONS should have certain information on file with local FCC offices, according to recently released FCC policy letter. This includes coded control signals for repeaters or telemetry data from a remotely controlled amateur station, for example. The required information must be filed with the Engineer-In-Charge of the FCC district from which the coded transmissions will be made. Compliance of the operators involved is requested at the "earliest possible convenience."

CURRENT FCC AMATEUR DOCKETS have as yet received no formal attention from the commissioners. Each docket requires a good deal of staff work prior to consideration in the form of evaluating the comments and reply comments and incorporating them into a proposed action. It looks like the first amateur docket to be put before the commissioners will be 20001, Commemorative Callsigns. Somewhat later will be 20092, Permitting Extra Class licensees to request specific callsigns. But then again, both could be held up along with 19658, License fees. Not only are various pending dockets inter-related, but the possible effects of "restructuring" must also be considered.

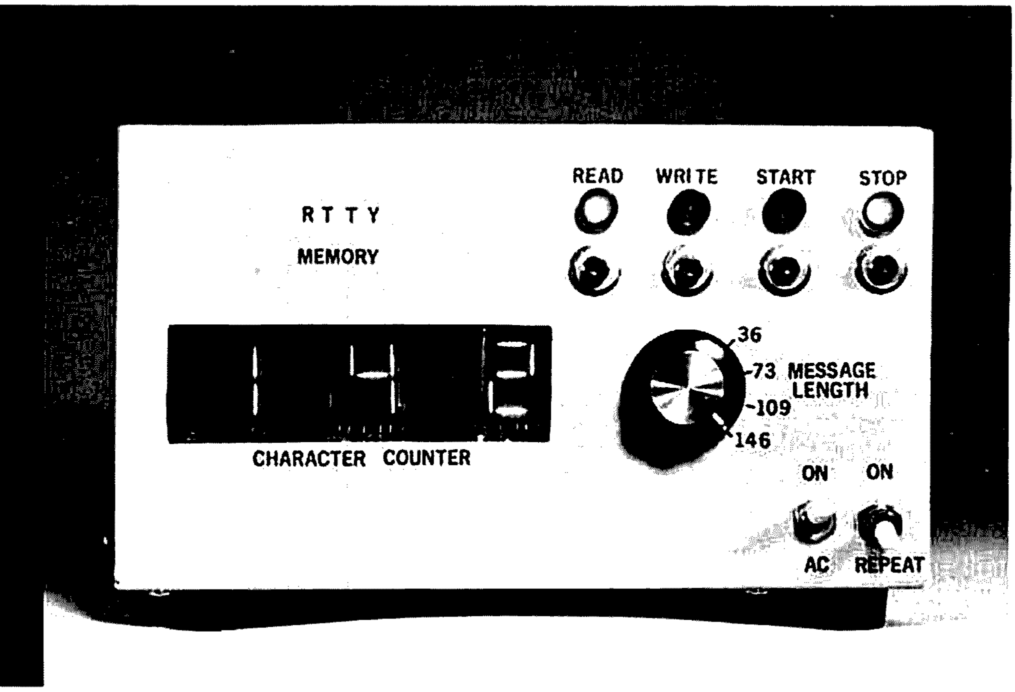
RECENT HOT 10-METER CONDITIONS have pointed up the value of beacons as propagation indicators. Gary Hendrickson, W3DTN, has been talking up establishment of two-meter beacons on some "little-used" FM simplex channels -- but finding such a channel nationally would be an almost impossible task.

Consider Adopting a "split-split" channel for this purpose, for example 146.655 or 146.625 MHz. Stock crystals could be pulled on-frequency in most receiver circuits, and with a very narrow deviation MCW ID every two or three minutes, little if any problem would be felt by nearby simplex activity.

Repeater Council and Frequency Coordinator Reactions to the idea are solicited; some very interesting and potentially valuable information could result from a comprehensive beacon net.

BEHIND-BARS HAM RADIO POSSIBLE, according to a reply by A. Prose Walker to a question posed by Mid-Oklahoma Repeater Inc. (MORI) officer. Idea behind the question was that amateur radio could be a powerful rehabilitation tool, both in prisons and in "halfway house" environments. Position of FCC is that conviction for a crime -- particularly if radio was not involved -- should not bar one from holding a license.

VLF BUFFS HAVE AN SWL GROUP -- Long Wave Club of America -- with much useful info for ham VLF experimenters. Club has a free monthly newsletter, Lowdown, and club President John Clements also offers a useful list of over 2500 VLF stations for sale at $2.00, postpaid. Contact John Clements, LWCA, 11425 Albers Street #5, North Hollywood, California 91601.



random access memory RTTY message generator

Construction of a solid-state memory that replaces the tape distributor and reperforator used in amateur RTTY stations

Bert Kelley, K4EEU 2307 South Clark Avenue, Tampa, Florida

This RTTY accessory uses Random-Access Memory (RAM) ICs in an easy-to-build, compact package that replaces both the tape reperforator and transmitting distributor for station call-ups, CQs or special applications such as RY generation. It is considerably easier to program than other devices. The message (data) is entered into four RAM chips using the station teleprinter keyboard – it can be recalled instantly at the touch of a button.

Operation is noiseless and there is, of course, no tape to handle. The memory will record (write) from a TD, off the air or the keyboard of any five-level teleprinter such as the model 15 or 28. As an RY generator, line feeds and carriage returns may be added to prevent end-of-line pile up, a convenience the usual RY generator lacks. Although this RTTY memory is designed to be used with ST-5 or ST-6 demodulators,[1] flexible input and output circuits permit use with other demodulators or simple teletype loops.

Random access memories using mos technology are available in surplus at quite reasonable prices. While other chips are available with large capacity memories, the 1101, a 256-bit chip, is selling at $2.56 at time of writing, or one cent a bit. The effective memory capacity of 1024 bits is obtained by ORing four

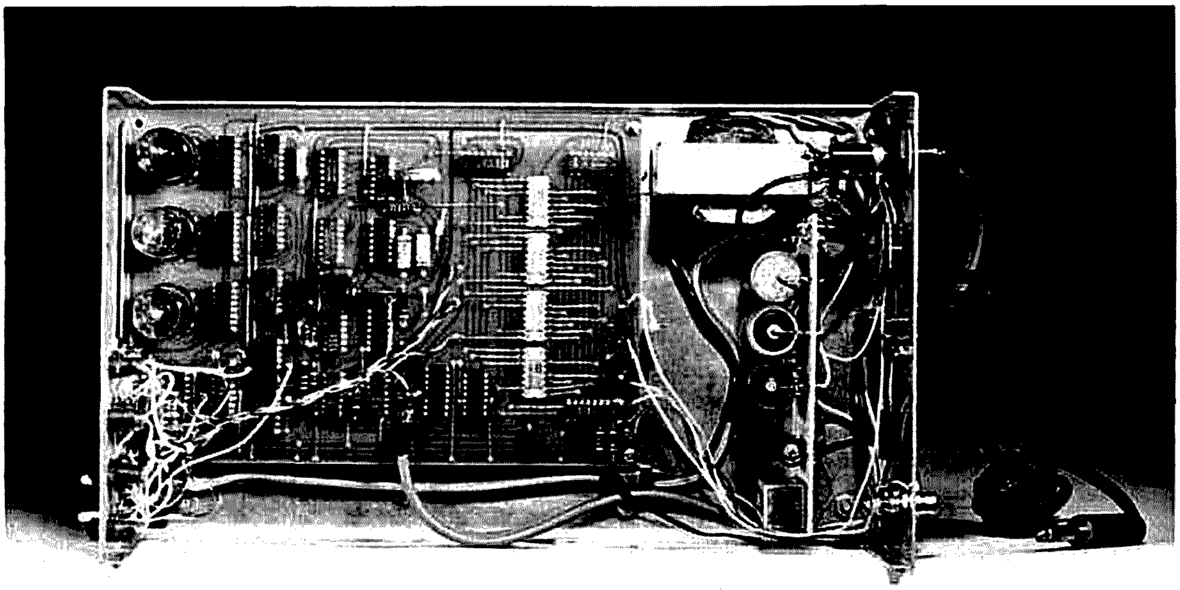
chips together and sequentially entering and retrieving the RTTY signal. The output is a regenerated, low-distortion signal, and playback proceeds smoothly as operation transfers from chip to chip.

operating principles

Each RTTY character at 60 wpm is composed of a 22-millisecond start pulse, five 22-ms coding pulses which determine the character, and a 31-ms stop pulse. These seven pulses are converted to TTL logic levels and serially entered into the RAMs at bit locations in the memories determined by the binary status of eight address lines. Data transfer is accomplished by a positive write pulse at the center of each 22-ms interval. Since a short time is specified[2,3] before the write pulse is applied after memory address, and because of marking and spacing bias present in many signals,[4] a short write pulse positioned at the center of each bit is most immune to these sources of distortion and provides best results.

Most of the ICs in the circuit are used to generate timing pulses and waveforms which coincide in time with the input signal from the keyboard during the write mode and cause the memories to read out an accurately-timed 60-wpm signal during read.

It is easy to generate the 22-ms intervals, but the odd 31-ms pulse poses a small problem. One solution is to generate the stop pulse as two 22-ms pulses. This will work, but the playback is slower than standard speed and there are now eight instead of seven bits to make up each character, reducing the total memory capacity. Also, it would not be possible to record a properly operating TD because the memory could not keep up with it.



Layout of the RTTY memory showing printed-circuit board and Minibox chassis. Relay and power-supply filter capacitors are mounted on piece of perforated board to the right.

I decided to generate the required 31-ms stop pulse by writing the 11-ms into the memory with the clock, then generate an additional 20-ms stop pulse with a one-shot multivibrator to stop the clock oscillator. For a character seven bits long, each 256-bit memory chip has a capacity of 36-4/7 characters. For example, a message that exceeds the capacity of the first memory transfers into the second memory at the 37th character by writing the start pulse and the first three coding bits in the first chip, and the last two coding bits and stop

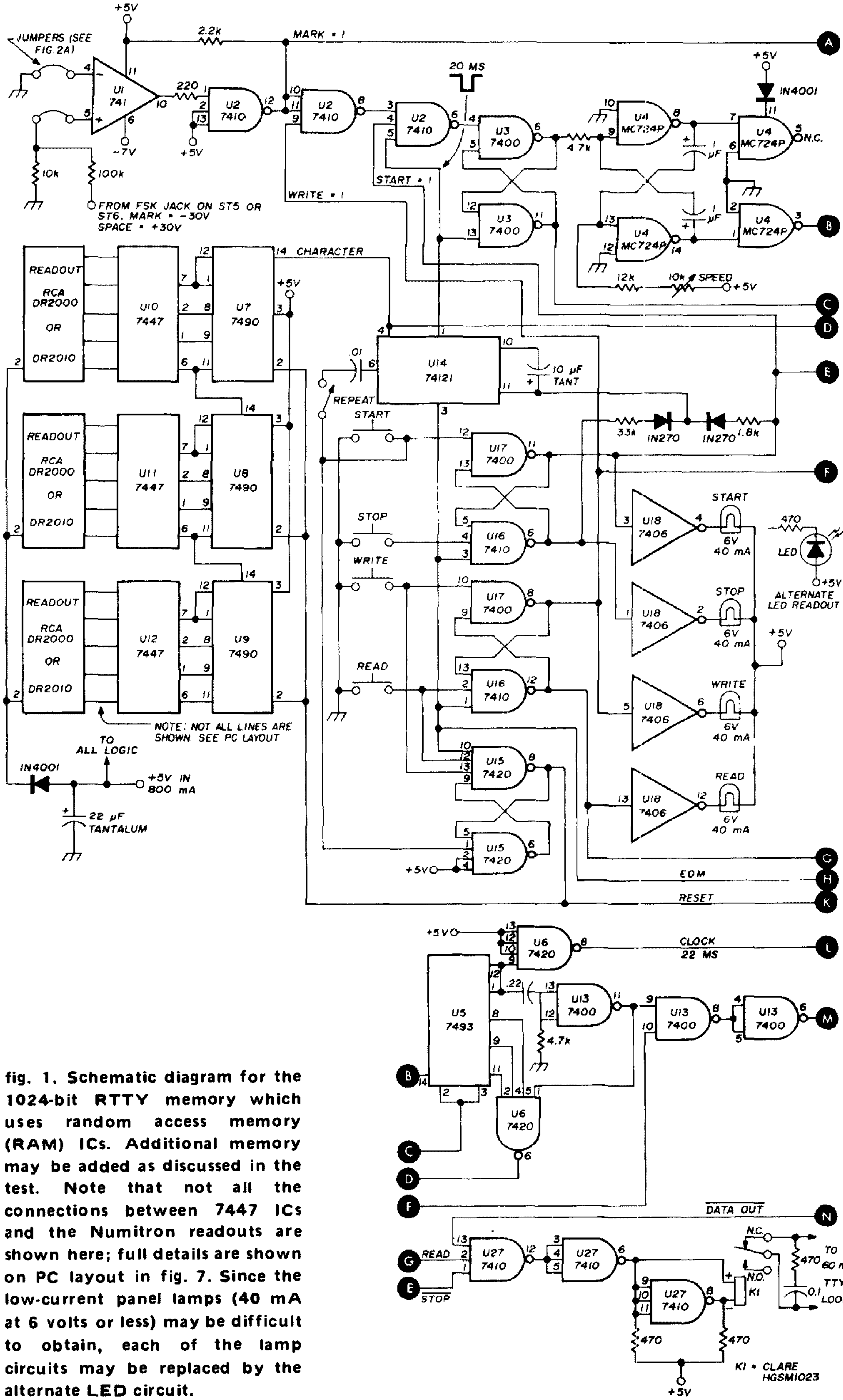
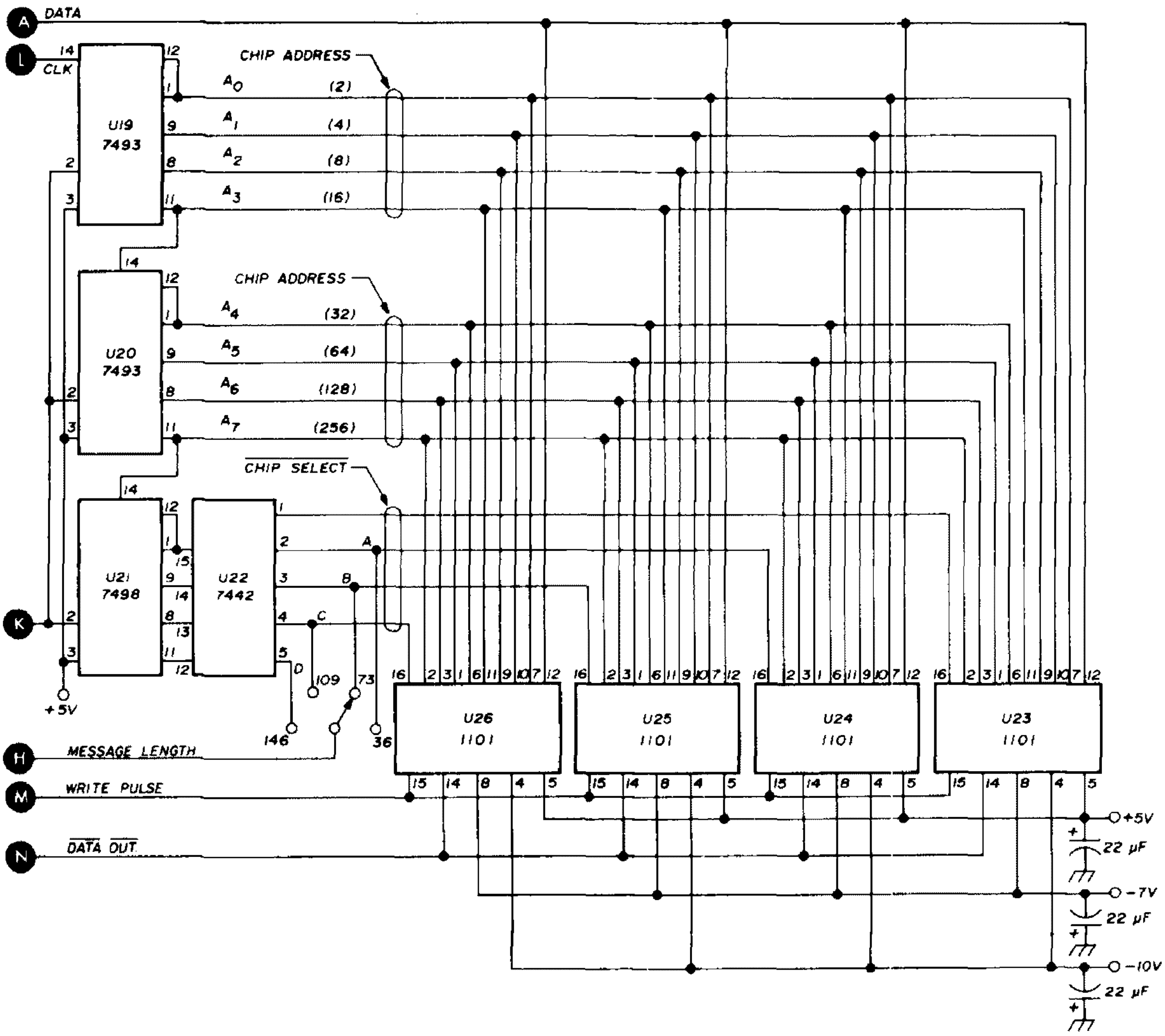


fig. 1. Schematic diagram for the 1024-bit RTTY memory which uses random access memory (RAM) ICs. Additional memory may be added as discussed in the test. Note that not all the connections between 7447 ICs and the Numitron readouts are shown here; full details are shown on PC layout in fig. 7. Since the low-current panel lamps (40 mA at 6 volts or less) may be difficult to obtain, each of the lamp circuits may be replaced by the alternate LED circuit.



pulse on the second chip. Additional circuits are provided to control the functions of read, write, stop, start and repeat. A three-digit character counter indicates the exact position of the memory in use.

circuits

A 741 operational amplifier is used in a versatile input circuit which can be used with positive/negative or positive-only keying voltages. This device has high gain and tolerates unbalanced supply voltages well. The differential input of the 741 amplifies the difference between the voltages on pins 4 and 5. The main requirement is that the input voltage swing should not exceed the supply voltage in either direction – the 100k resistor should be changed as necessary to comply with this requirement.

The input connections to pins 4 and 5 should be made so that the output of U1 (pin 10) pulls U2 (pin 1) to logic zero during mark, and logic 1 on space. The FSK keying voltage of the ST-5/6 swings both positive and negative (referenced to ground). Therefore, if the input is connected to U1 (pin 5) and U1 (pin 4) is grounded, proper drive results at the TTL gate.

Fig. 2A shows pin 4 of U1 biased above ground so positive-only keying voltage will result in a negative output at U1 (pin 10) during mark-hold. An FSK voltage that is positive during mark-hold should be connected to the inverting input of U1 (pin 4) through a voltage divider. TTL nand gates have built-in diodes to limit voltage reversals at the input, and the resistor limits the diode current.

The RTTY signal is translated to TTL

logic levels at U2 (pin 12). The next section of U2 acts as a gate to prevent incoming signals from affecting the clock when in the read mode. When recording into the memory, both U2 (pin 9) and U2 (pin 4) are high and the incoming start pulse from the keyboard goes to logic zero at U2 (pin 6), setting the R/S flip-flop U3. U3 (pin 6) goes to logic 1, the RTL clock multivibrator U4 immediately starts oscillation at 91 Hz and U5 divides this to a 22-ms square wave which drives the address counters U19 and U20 through U6 (pin 8).

The negative-going center of the symmetrical square wave at U5 (pin 12) is used to generate a positive pulse at U13 (pin 11). The other section of U6 generates a negative-going pulse (at U6, pin 6) when U6, pins 1, 2, 4 and 5 are all high at the same time. This occurs eleven milliseconds into the seventh bit, i.e., the stop pulse. This pulse then triggers U14 (pin

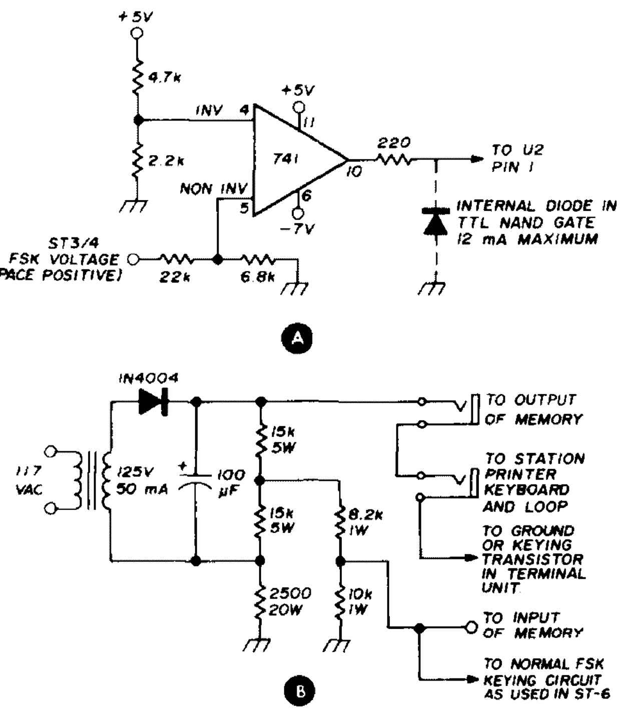


fig. 2. Circuit in (A) shows how op amp input circuit can be changed for positive-only keying voltage such as the ST-3 or ST-4. Printed-circuit connections to pins 4 and 5 of the 741 are designed so they can be jumpered for the desired input configuration. Circuit in (B) shows how RTTY memory can be connected to ST-6 terminal unit, or a circuit built for local loop operation. The input to the memory does not load the ST-6 and affect the normal FSK connection at the same point.

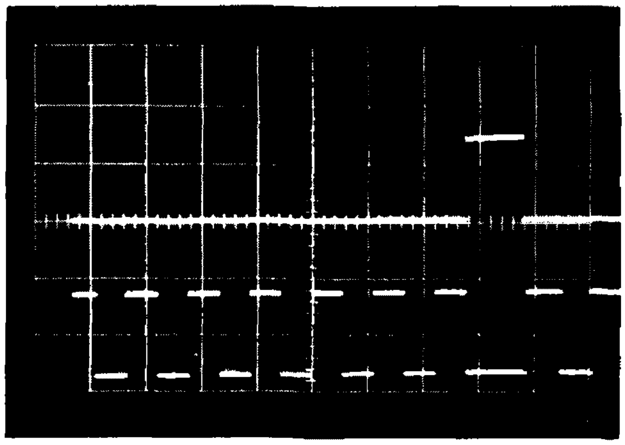


fig. 3. The top trace shows the single output pulse at U14, pin 6. The bottom trace shows the clock signal at U6, pin 8. Horizontal scale is 20 ms per division; vertical scale is 2 volts per division.

4), the monostable miltivibrator, which stops the clock by generating a 20-ms pulse which resets U3. The scope traces (**fig. 3**) show the pulse-timing relationship between the clock, write pulses and the output of U14.

U13 also generates a positive pulse train during the write mode only (**fig. 4**). U19 and U20 generate the binary address voltages for the RAMs which are applied to all four memorys. The RAMs are three-state devices that have a logic, zero, logic one and a third inactive state which is maintained as long as there is a logic 1 on the chip select line, pin 16. As each RAM is addressed at all 256 locations, the voltage on U20 (pin 11) falls, advancing ICs U21 and U22 ahead to select the next memory chip, or activate the End of Message (EOM) circuit.

The EOM pulse stops playback or recording, resets the character counter to 000, resets all address counters, and triggers U14. Pulse lengths from U14 are determined by the status of the start/stop latch, and U17 (pin 11) is normally high during operation so U14 generates the 20-ms pulse. However, U16 (pin 6) goes high when playback stops and a long stop pulse is generated because the 33k resistor now sets the timing. This allows partially entered characters at EOM to clear the page printer before the new readout and another start pulse.

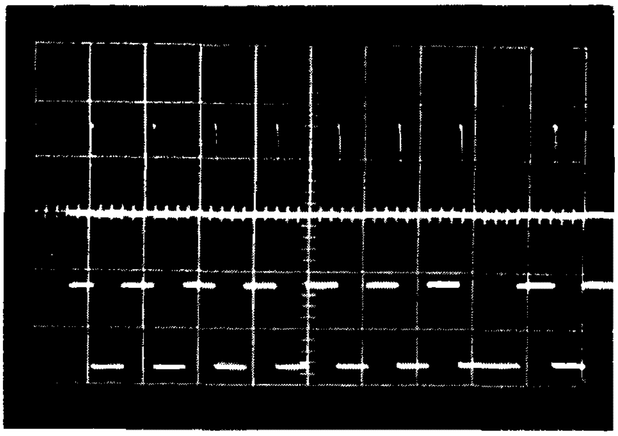


fig. 4. Narrow pulses on upper trace are output at U13, pin 11, during read and write (same output at U13, pin 6, during write only). Bottom trace shows clock signal at U6, pin 8. Horizontal scale is 20 ms per division; vertical scale is 2 volts per division.

At EOM the sequence is as follows: generation of a long stop pulse at U14, momentarily readout stop, then the start latch is re-triggered by the trailing edge of the pulse from U14 (pin 6). During read the clock is gated by the output of U14, gate U27 is enabled, and the data output operates the high-speed keying relay. During write, U27 is placed in mark-hold, keeping the loop closed.

The TTL logic and Numitrons are operated from the same 5-volt power supply (fig. 5). The readout voltage is reduced by using the fixed voltage drop across a silicon diode since the full 5 volts is not needed. The LM309K regulator should be mounted on a heatsink with a small amount of silicon grease applied to help heat transfer. Measure the supply voltage under load and reject the regulator if it does not supply 4.8 to 5.2 volts. It appears that some devices sold as LM309Ks will not meet this test.

Two negative supplies are shown for the mos devices because the Intel 1101 RAMs I used arrived with a spec sheet showing −7 volts V_{dd}, and −10 volts for V_d. The Texas Instruments TMS 1101NC is similar but requires −10 volts on both pins 4 and 8, and the Signetics 2501 calls for −9 volts. Since you may have different devices available, it is a simple matter to change the power-supply voltage by installing a different zener diode.

In the process of testing this unit, memory ICs were obtained from different supply houses. Many were unmarked, others had different numbers on them. Results of tests showed most devices operated better with a single negative

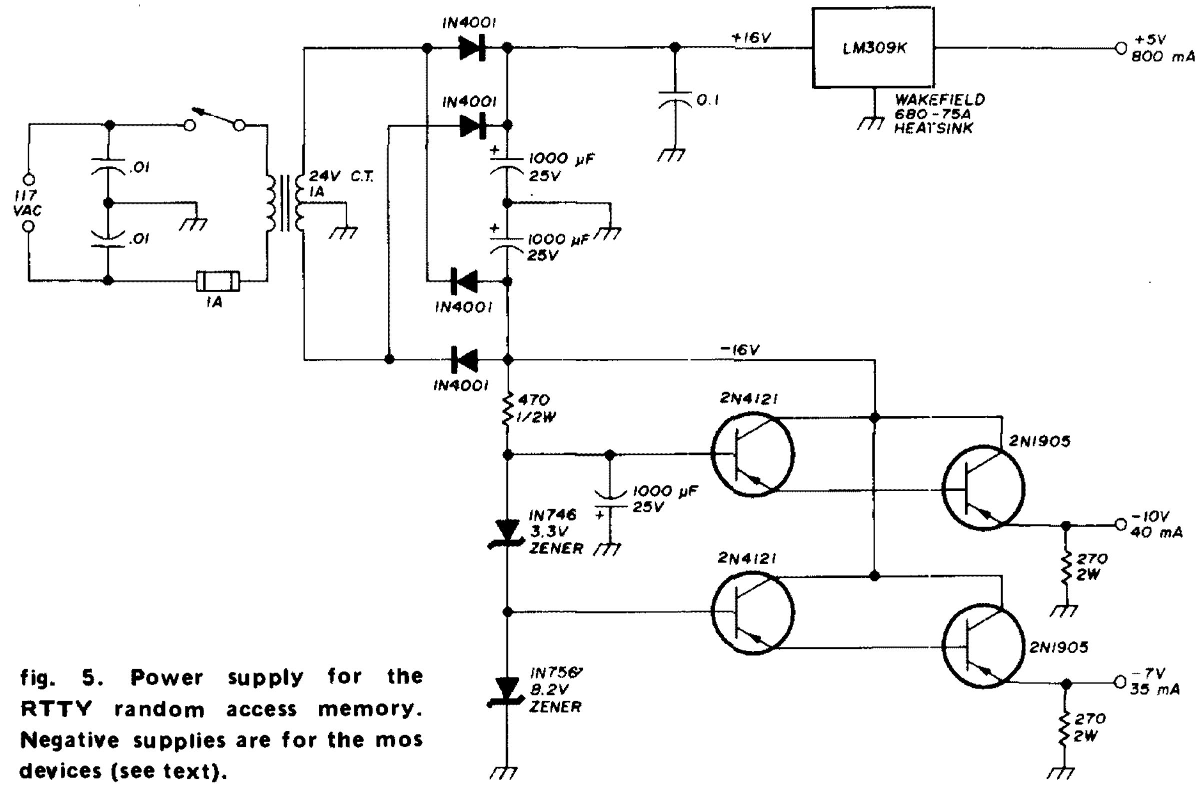


fig. 5. Power supply for the RTTY random access memory. Negative supplies are for the mos devices (see text).

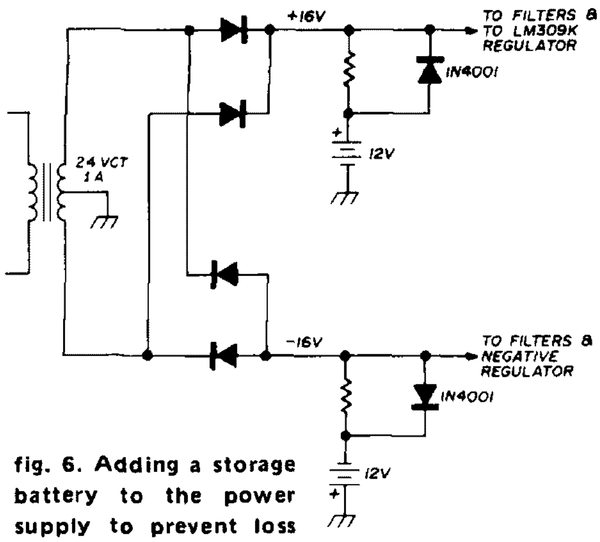


fig. 6. Adding a storage battery to the power supply to prevent loss of programming due to momentary power loss.

supply voltage of from −7 to −10 volts for both V_d and V_{dd}. It also indicated the advisability of using IC sockets for U23, U24, U25 and U26 (installed after the photographs were taken).

The power-supply filters and keying relay are mounted on perforated circuit board. The relay used in this model is a CP Clare HGSM 1023, a high-speed, mercury-wetted, bistable, single-coil relay. It has a low power requirement and operates by reversing the current through the coil. If this relay cannot be obtained, the Clare HGS 5006 or P&B JMF 5080 appear to be suitable substitutes. U27 could also drive a high-voltage transistor such as the HEP244 to key a loop directly or to operate a higher voltage, fast acting, mercury-wetted relay.

The device requires so little power that it is practical to leave it on continuously and retain, indefinitely, anything programmed into the memory. **Fig. 6** shows how two small storage batteries will make the RAMs immune to momentary loss of power. The resistors are selected to keep the batteries on trickle charge. When ac power fails the diodes bypass the resistor and immediately assume the load.

construction

Most of the construction work is already done for you if you purchase the circuit board,* A layout is given in **fig. 7** for those readers interested in making their own etched boards.[5] A photo shows jumpers required but does not show those located under U7, U8, or U9 – check before installing the ICs. Parts values and all IC numbers are screened on the 5½x8-inch (14x20.3-cm) circuit board where space will permit. The assembled board is mounted on spacers in a 12x7x4-inch (30.5x17.8x10.2-cm) Minibox. The end-of-message switch, pushbuttons and toggle switches are small imports available from Lafayette or similar sources. Panel lamps should be low current, 40 mA or less at 6 volts, to stay within the rating of the 7406.

The installation of the mos RAMs should be saved for last since, like mosfets, they can be damaged by static voltages.[6] One manufacturer recommends

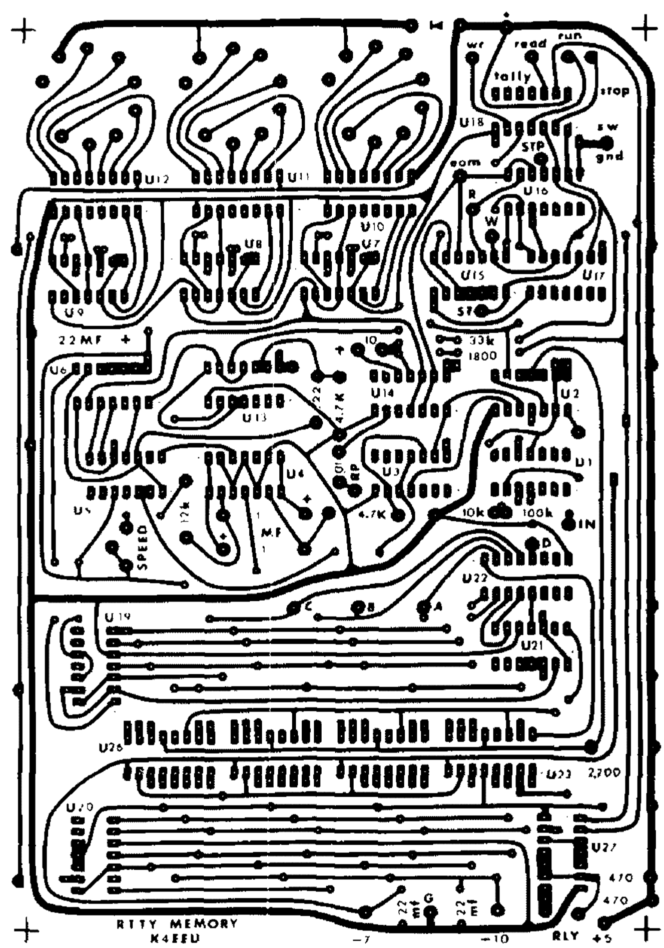


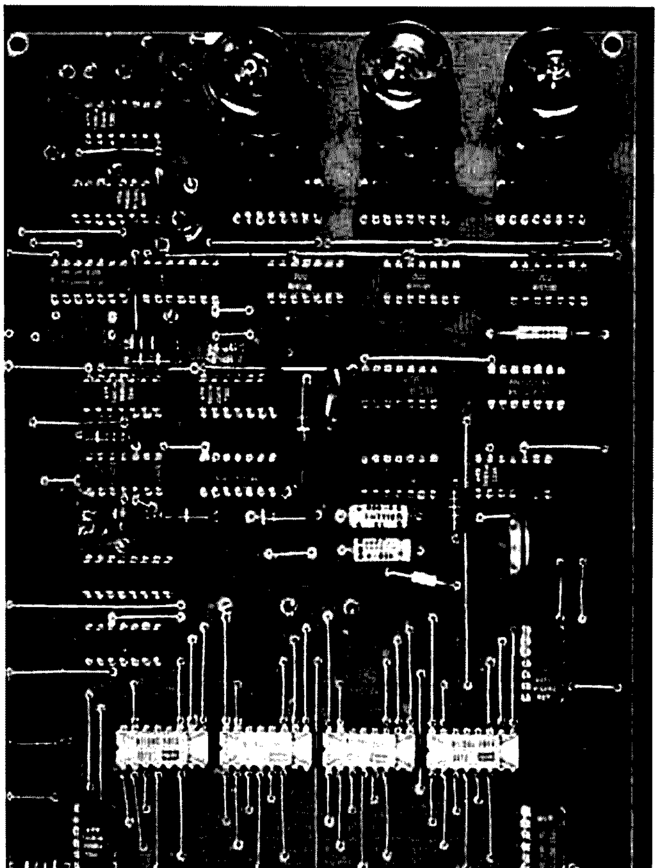
fig. 7. PC layout (shown half size) for the RTTY memory is designed for DIP-packaged ICs. Numitron readouts plug into CJ9PCB1 sockets which are soldered to the circuit board.

*A plated, undrilled, epoxy-glass printed-circuit board with enlarged assembly photograph is available from the author for $10, postpaid in the United States and Canada.

grounding the soldering iron and advises against bending leads close to the IC package. The possibility of damage is reduced considerably if sockets are used and all power-supply voltages checked before connecting the board.

checkout

When the unit is turned on, the read or write pushbuttons should set the character readout to 000, and the teletype machine should not run open since U27 (pin 12) should be at logic one, or mark-hold. It may be necessary to reverse the coil leads of the relay. Start the memory in the read mode and set the period of the square wave at U19 (pin 14) to 22 milliseconds and confirm that the stop pulse at U14 (pin 6) is 20-ms long. The character counter should be operating and the printer will print garble, if turned on.



Component layout of the printed-circuit board. Most capacitors are low-voltage tantalum units. Numitron readout tubes are plugged into sockets. Printed-circuit layout is shown in fig. 7.

With the input of the memory connected to a suitable FSK or keyboard output, type a short test into the unit and test the playback. The character counter should follow both the entry of the test into the memory on write, and the playback. If all is well, perfect playback will result. However, it's likely that there will be some small problems.

It is important that the clock be operating at the right speed. If the device prints a few errors, this should be checked. Note where consistent errors occur in the print, if at the same position. This may be caused by a defective memory chip. It is not unusual for a low-priced surplus memory to have two or three defective cells out of the possible 256. These will cause substitution of a different letter from the one entered in the write mode. A five-level coding chart can be used to determine what is happening, for those interested. A faster and more practical solution is to plug in a spare memory chip. Also, don't overlook the possibility of a defective input signal from a noisy printer keyboard.

general

Construction cost should be about $80. This can be reduced, at some loss in operating convenience, by omitting the character counter. The memory capacity can be more than doubled by adding five more RAM chips on an external board, running parallel address lines, and using the spare outputs of U22 for chip select lines.

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ham radio

low-cost printed-circuit boards

A simple technique for building low-cost printed-circuit boards for your homebrew amateur equipment

Bill Wildenhein, W8YFB, 41230 Butternut Ridge, Elyria, Ohio 44035

Many older amateurs and a surprising number of beginners are reluctant to try building their own printed-circuit boards. Many feel that PC boards are too inflexible to permit circuit changes or the cost of materials and equipment is high. Others think PC board layouts require special knowledge, skills and equipment.

Let's consider the first charge: Inflexibility. If you make a complete receiver layout on one board it will certainly be inflexible! Look at **fig. 1**. That curious mess is actually a pretty hot 20-meter receiver. Notice that it consists of a number of small modules. Most important, it will perform beautifully just as illustrated. In short, stop thinking in terms of wiring components together, and think about wiring complete stages together.

With this technique you can build progressively, and need not invest in high-cost items such as crystal-lattice filters until you are satisfied that performance is what you desire. Obviously, this is a good way to avoid mistakes in planning chassis layouts. With a complete receiver running on the bench you will be able to envision layouts that will eliminate the need for conventional chassis design.

An example of this is shown in **fig. 2**. This ssb receiver was originally built as in **fig. 1**. Instead of buying an expensive enclosure, it was housed, as shown, in bent-up aluminum scrap with a steel wraparound. When I find a reliable and stable circuit and layout, I usually etch a number of boards "for stock." This results in terrific time savings later. For example, I wanted to experiment with a variation of the KY crystal filter which is popular in Europe but had no receiver to match the available crystals. With all but one board in stock it took only two evenings to build the receiver shown in **fig. 1**.

With this building technique, if you don't need the receiver, or tire of it and want to make something more sophisticated, your loss is zero. Simply disconnect the boards and put them away, intact, for the next project. Another advantage is this: Many circuit boards can be designed as multipurpose boards. Certain transistors and ICs, for example, can be arranged as rf and i-f amplifiers, mixers, product detectors or audio amplifiers. As will be shown later, the circuit board can also be devised to accommodate quite a variety of circuit variations and component selections.

Is the cost really high? The best circuit-board material is the relatively expensive G-10 fiberglass-epoxy material. This runs about $2.00 per square foot (929 square cm) from surplus houses such as Jefftronics, Meshna and others. The "universal" circuit board which I use

(described later) requires 2.35 square inches (15.2 square cm) of material. With the G10 material at 1.4 cents per square inch (6.5 square cm) the universal board will cost 3.4 cents. At that price it hardly pays to economize by using the less expensive phenolic board.

etching circuit boards

The etching supplies and equipment you will need are shown in fig. 3. The two-pound coffee can houses a 40-watt light bulb. This is a heater for the etchant, and a stand to hold the plastic cereal bowl firmly while etching. Etchants are very corrosive, so a spill can be messy. The household cleanser is needed for scrubbing boards, and the masking tape is about the cheapest and most convenient resist available. The plastic funnel is used to pour the etchant back into its container.

All of your etchant-handling equipment should be plastic. If you have had little or no etching experience, buy a prepared solution for your initial work. It will give you a feel for the proper mixture and proper etching action. Two widely available sources include Radio Shack and any dealer handling the popular GC (General Cement) line of radio and TV service products.

The GC etchant is about the highest priced at six fluid ounces (177 cc) for $1.20, so let's use it as an example. That bottle will etch about 100 square inches (645 square cm) of copper. The universal board to be described requires only 0.85 square inches (5.5 square cm) of the total 2.35 square inches (15.2 square cm) to be etched away. This is typical. Including waste it comes to about one-third the total board area. Thus, the little bottle will actually handle about 300 square inches (1935 square cm) of board material, or about 0.4 cent per square inch (6.45 square cm). The universal board will require about 1-cent to etch, making total board cost about 5 cents. This is pretty inexpensive for an entire stage.

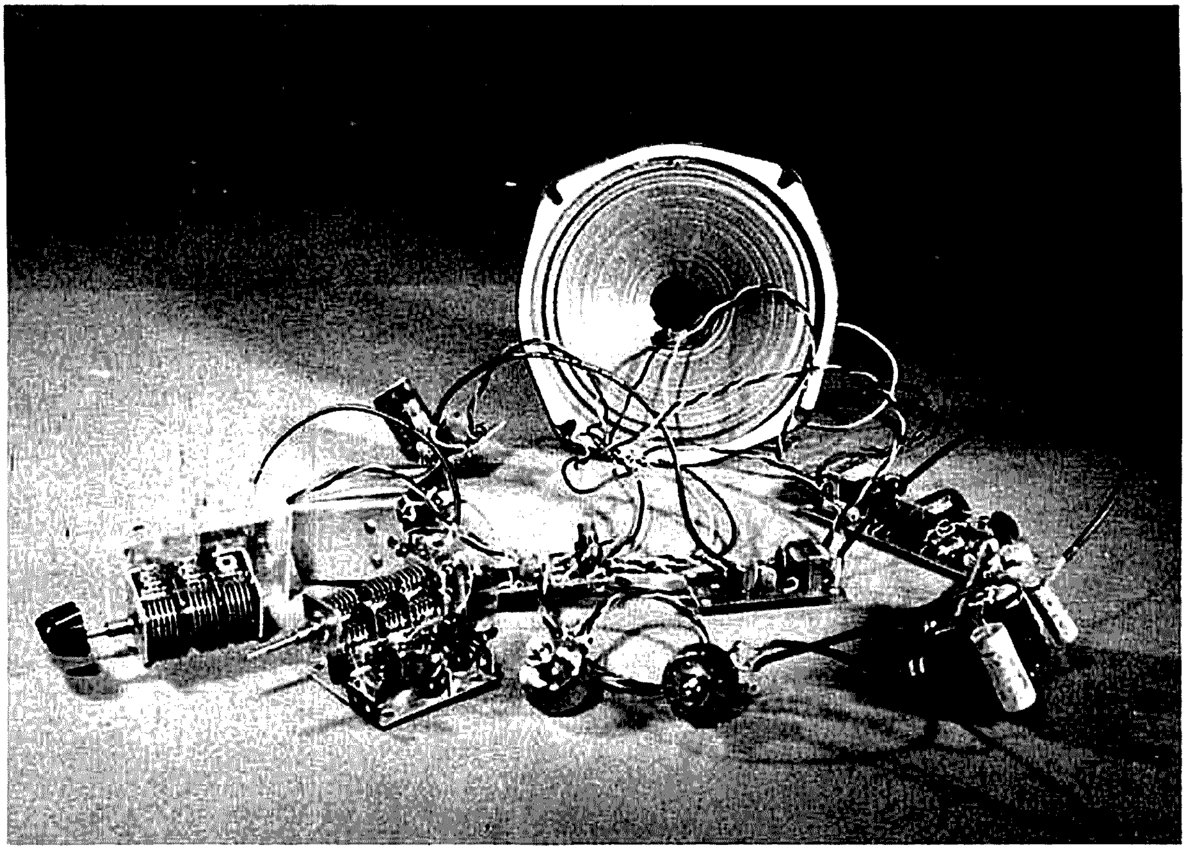


fig. 1. Twenty-meter receiver built with low-cost printed-circuit boards. Careful attention to circuit design provides high performance, even in this haywire condition (see text).

Be sure to select ferric chloride for your etchant. Ammonium persulphate is also available. It works well with photoresist techniques, but with our cheap masking-tape resist, it will do a ragged job, tending to etch under the masking tape. Ferric chloride will do a clean, neat job everytime – even with an old, loaded solution.

board production is more a matter of common sense than specialized skills or knowledge. I usually start with a schematic of a board, and try to include any expected variations to make the board fit different components and circuit variations. For the sample board I will use four circuits that are similar. All use a low-cost dual-gate, gate-protected mosfet.

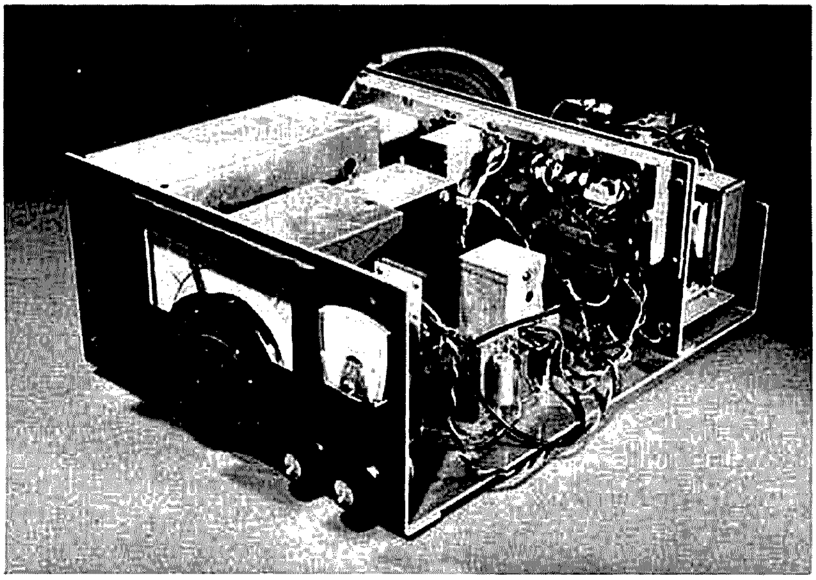


fig. 2. Packaged single-sideband receiver using the components shown in fig. 1.

Two items not shown in the photograph are the drills you will need, and a pad of cross-ruled paper or graph paper (available in most discount stores). The drills most needed are numbers 60 and 54. If you are in an area where numbered drills are not available, you can use 1/32-inch (0.8-mm) and 3/64-inch (1.2-mm) drills. If possible, use industrial quality drills. Cost is about a dime more, but the life of the drills is often twice or three times that of discount store grades. If you have a choice, use high speed steel instead of high carbon. Many of the imported drills are just case-hardened, cold-rolled steel, poorly formed, and a poor investment.

The cross-ruled paper is a great help in making up layouts, and for maintaining a permanent record of your boards. From the foregoing it is obvious that the cost of materials and equipment is small.

circuits

As you will see, the layout work and

These devices aren't as fussy as the older non-protected mosfets and can be handled as easily as conventional transistors.

Fig. 4 shows the circuit for a mixer stage. This circuit is shown connected to a hot tank circuit where you must isolate the +12 volt supply from the gate. Where the tank circuit is at ground potential, C1 and R1 are unnecessary. The output is tuned with a small compression trimmer capacitor. Two types of output coupling are also shown. For convenience I usually use miniature coax (RG-174/U) to bring the vfo signal to the mixer board.

Notice that the +12 volt line is decoupled with R4, C5 and C6. This extra precaution results in greater stability, and is one of the things that provides stability in a haywire mess like that in fig. 1.

The circuit of fig. 5 can be used as an rf or i-f stage. Due to the extremely high gate impedance, feedback can be troublesome when high-Q tuned circuits are used. An i-f amplifier usually depends on the first filter for its selectivity so you

can afford to swamp the input with a 2200- or 2700-ohm resistor. This will make a completely stable system.

The receiver in fig. 2 uses two of these stages. The first is swamped with the filter terminating resistor, the second stage with 2200 ohms – the two stages become very docile. As an rf amplifier the circuit will require careful shielding of input and output circuits. By reducing the value of C5 and adding a "gimmick" capacitor or small adjustable capacitance between the junction of C5, R4 and L2 to gate 1 you can include neutralization with the same layout. Agc or mvc can be included as shown. Limit this voltage to ±4 volts on gate 2. Maximum gain requires about +4 volts from gate 2 to the source.

Fig. 5 includes another form of output circuit. The slug-tuned coil can be a TV i-f coil rewound for your frequency. For this service you must provide space to drill a 5/16-inch (8-mm) hole to mount the coil. If you don't need a low-impedance output, capacitor C9 can be connected across the coil and C10 can be eliminated.

Fig. 6 shows the circuit as a high-gain product detector. This circuit has good dynamic range and a microvolt input will deliver about a millivolt output. Notice that this circuit is identical to fig. 4 except for the details of the output circuit. The output transformer is a regular Radio Shack item. The 500-ohm output is convenient if you go into a low-pass audio filter built from 88-mH toroids, as it provides a good impedance match.

Notice the volume control hookup. If the volume control is located remotely it is not desirable to ground the low side of the control to the chassis at the remote point. This can lead to ground loops. Instead, carry the ground to the control with the shield of the miniature coax, and from the control to the next stage on the shield of that same piece of coax. This ground loop problem is very easy to control in a modular setup like this. You can mount the boards on pieces of bakelite or plastic when necessary and have full control of the ground situation without any radical rebuild.

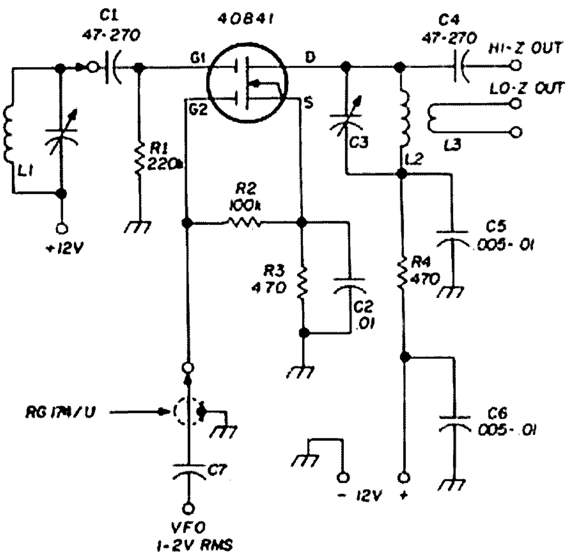


fig. 4. Mosfet mixer stage designed for printed-circuit use.

fig. 3. Etching supplies and equipment you need for making your own low-cost printed-circuit boards.

Notice that the output is bypassed for both rf and audio (C5 and C10). Also notice C9. This capacitor can roll off the high-frequency audio components. That little transformer is quite efficient out to 70 kHz or more, and there is no need for response above about 2.5 kHz. In fact, if you use a 0.05-μF capacitor at C9, you will have a definite peak around 1 kHz that is helpful for CW work, but still results in useable ssb performance. Again, the bfo voltage is brought in with RG-

174/U coax. With a high-gain i-f strip, it may be helpful to locate the bfo well away from the strip – or shield it.

The receiver in **fig. 1** shows some i-f overload, but by tucking the bfo module behind the speaker, operation is completely satisfactory. Slightly more gain will result if C2 is paralleled with an

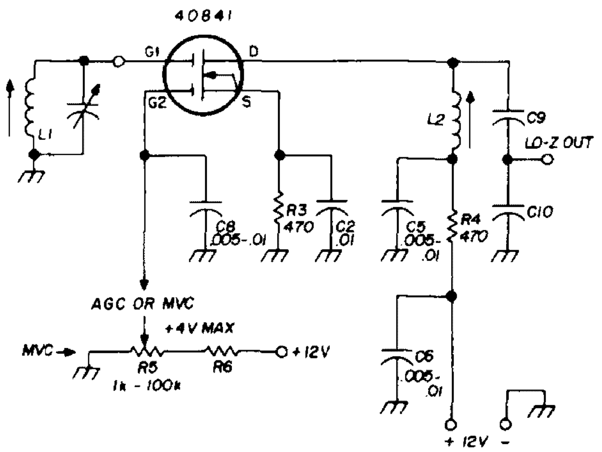


fig. 5. This circuit can be used as an rf or i-f amplifier and uses same printed-circuit layout as the circuit of fig. 4.

audio bypass capacitor, although it might be more worthwhile to go to a Motorola MC1550 (HEP 590) or RCA 3028 IC. They will deliver considerably greater gain if you need it.

The circuit in **fig.** 6 can also be used as a *synchrodyne* detector. Tune L1 to the operating frequency, feed in a vfo signal at the operating frequency, add an audio filter to the output and use a high-gain audio amplifier. You now have a simple, direct-conversion receiver.

Fig. 7 shows the circuit for a grounded-gate rf stage. Note that gate 2 must still be biased to +4 volts for full gain. You can feed gate 2 from the agc line if desired. This stage is very stable without much shielding.

board layout

If you can lay out a single circuit board to accommodate these four circuits you will have a really useful board! **Fig.** 8 is the result. The simple square layout makes it easy to cut out. Notice the fat ground bus. This is a big help in achieving stability in rf equipment. In my layouts I usually make any unused board space a part of the ground bus, rather than etch it away. You save etchant and gain stability.

Circle B is the location for a 5/16-inch (8-mm) hole for TV-type slug-tuned coils, or a 3/8-inch (9.5-mm) hole to clear the mounting tabs on compression trimmers. If you allow the tabs to protrude through the board at this point, it's advisable to scrape away the copper foil so the tabs are not grounded. Just a tiny chamfer is sufficient.

Locations A and C are for the leads of compression trimmers. These are usually 25/32-inch (20-mm) apart. By relocating A and C to the right so their center line lies 9/16-inch (14-mm) from the right edge of the board instead of 3/4 inch (19 mm), and drilling them 7/8-inch (22-mm) apart instead of 25/32 inch (20 mm) you can accommodate the small surplus Erie ceramic trimmers. If you make up boards in advance these holes can be left undrilled so that the board can fit any combination of output parts later.

Fig. 9 shows the mounting of the little Radio Shack transformers. The mounting tabs are pushed through the holes and bent over. It's best to solder one or both tabs to the ground foil. This will result in a quieter circuit. Use 4-40 screws to mount the boards. Spacers can be scrounged from broken rotary switches, switching TV tuners, 1/4-inch (6.5-mm) OD steel or copper tubing, etc. Keep the input and output sections apart when laying out the boards. If you lay out a

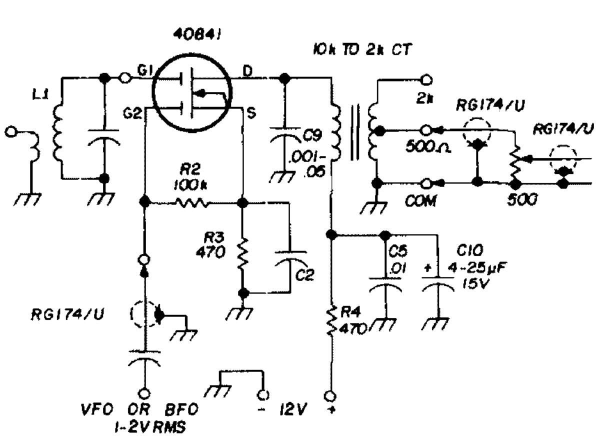


fig. 6. High-gain product detector. Except for the output this circuit is identical to fig. 4.

multi-stage board, don't use a U-shaped layout! This is asking for trouble. Keep the circuit in a straight line from input to output.

It's best first to plan your layout on the cross-ruled paper. Have typical parts available so you can allow for correct clearances. Once you have your final plan, keep the drawing for future reference. You would be amazed how fast you can forget what a particular board is used for, or where a particular part mounts. I keep a copy of the board, layout, schematics, and notes on performance or unusual characteristics together in a loose-leaf binder. That was one reason I could build a complete receiver in a couple of evenings.

When working with the TO-5 can integrated circuits, it is desirable to mount them for removal without damage. **Fig. 10** shows a method I use. Drill a 7/16-inch (11-mm) hole through the board, bend the leads as shown, and the IC can be soldered in and removed rapidly without damage.

circuit board fabrication

Now let's get to the actual board work. First, cut the board to size. A fine tooth hacksaw is best. File the edges smooth. Now thoroughly clean the copper side. Copper cleaner, other household cleansers or steel wool are all effective. It is very important to prevent fingerprints on the cleaned board surface. Often the film from a fingerprint seriously retards etching.

Next, cover the copper foil with masking tape. Choose a width of tape that will completely cover the board, if possible. Where tape is lapped, etchant often creeps under the tape. Press the tape down thoroughly all over the board. Draw the complete layout on the masking tape using a soft pencil. Be sure to indicate points to be drilled (I draw a tiny cross at these points). Cut through the tape with a very sharp jack-knife or X-acto modelmakers' blade. Peel the tape off from areas to be etched. Here, too, avoid getting your fingers on the exposed copper. Finally, when all the areas to be etched are exposed, again firmly press down the remaining tape with a clean pencil eraser.

Now, place the board in the cereal bowl, foil side up, and pour in enough etchant to cover the board to a depth of 1/4 inch (6.5 mm) or more. Place the bowl on your "heater" and relax. After

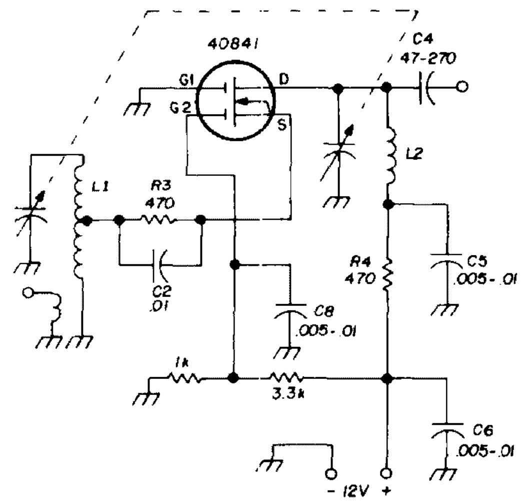


fig. 7. Grounded-gate rf stage.

about five minutes lift one end of the board out of the etchant using the plastic stick. As you lift it, you will see a dirty, dark residue draining from the board. Lift the end of the board in and out of the solution several more times to flush away this spent etchant, then lean back for several minutes more. Again flush away the loaded etchant. In about 20 minutes the board should be completely etched.

You can now remove the board from the etchant, rinse it in water (preferably not in your wifes' stationary tubs because this stuff stains badly). If you object to stained fingers, or have sensitive skin, use a plastic glove to handle the board.

Carefully check the board to make sure it is completely etched. If not, dump it back into the etchant. When it is completely etched, rinse it in water and damp-dry it with a clean cloth. Before removing the tape, drill the holes. Now you can peel off the tape and thoroughly clean the board with household cleanser to remove any vestige of etchant. If you are sloppy in this respect, a slight residue

of etchant can attack electronic components later.

Be very careful when handling the etchant! A tiny drop on a good tool will cause the tool to rust almost before you can blink your eyes! An accidental spill on the workbench will create an area that is saturated with etchant, and any tool or part that you leave on that area will be attacked – even months later! I use a little homemade stool in the middle of the basement floor for etching. When etching is completed, don't throw away the solution. Return it to its bottle. Use a plastic funnel.

After etching many boards you will notice that the etchant seems to be thicker, and it will be difficult to see the flushing action as you lift one end of the board. You may suspect the etchant is no longer usable, but a certain amount of water is vaporized as you heat the solution so if this condition appears, gingerly add water – usually a very small amount is necessary – to restore the solution to its original state. You will again be able to see the flushing action clearly, and etching speed will be better.

As you use your original bottle you will develop a good feel for proper consistency and etching action. Finally you'll reach a point where a properly thinned solution requires 45 minutes to an hour to do one board. Your masking tape will be badly discolored and it will be very difficult to see your drill location marks. By now you will be a pro with the stuff. You can buy the ferric chloride powder much cheaper, dissolve it in water, and be sure of your mix. Don't be surprised if you find others who are willing to pay for your services. Take the time to learn to do a craftsmanlike job and if your radio budget is low, you can probably stretch it a long way by doing circuit boards for other amateurs.

drilling boards

The drills for circuit boards are very small, but can be used in any kind of a drill motor. I have successfully used inexpensive hand drills (the eggbeater kind). If you buy one, take a 1/32-inch (0.8-mm) or number-60 drill along to be sure that the chuck will close on a drill this small. Select one that turns smoothly. A drop of oil on the gears and bearings often does wonders. When using such a drill, chuck it so only a short section protrudes from the chuck. Rotate the chuck smoothly and evenly, and don't try to rush the job or you'll break the drill.

If you use a drill motor, it is best to use fairly slow speeds. It may be worthwhile to make up an SCR speed control. Again, allow as little as possible of the drill to protrude from the chuck. Of course, if you have a drill press, there is very little trouble. When drilling circuit boards, however, you will find that drills do not last as long as they do when

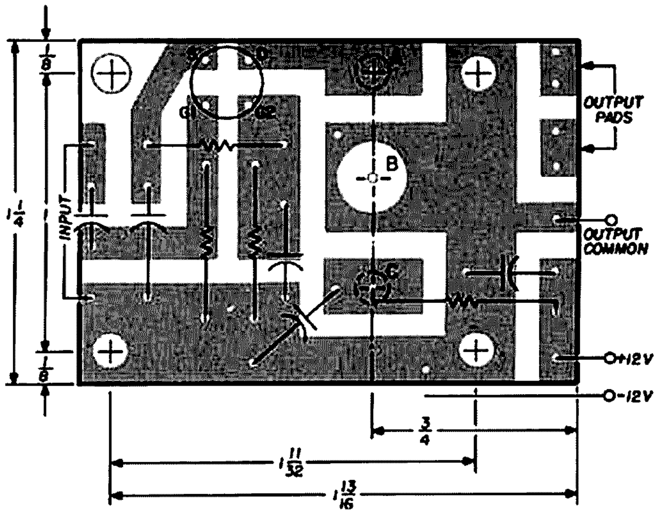


fig. 8. Printed-circuit layout which will accommodate the circuits of fig. 4, 5, 6 or 7. Mounting of the transformer for the product detector is shown in fig. 9.

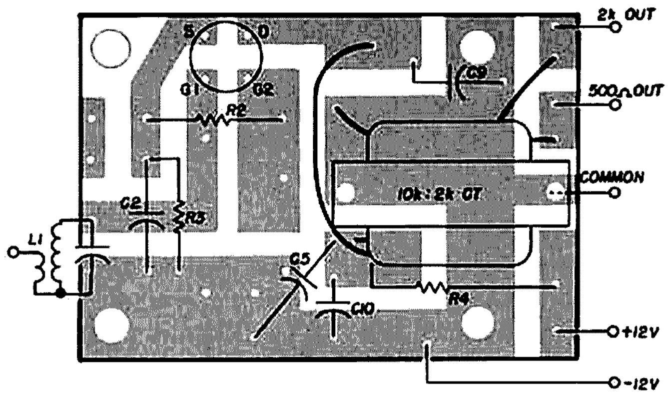


fig. 9. Method of mounting the product-detector output transformer on the universal printed-circuit board.

drilling steel. Fiberglass and phenolics are both abrasive materials. You may get only 100 holes with good grade drills, then the drill begins to turn up an excessive burr (see fig. 11). This makes soldering difficult because the solder doesn't want to climb the burr. Also, when a drill gets to that stage, it will often wobble as it cuts, producing an oversize hole.

If you want to take a whack at this sort of construction, but are unsure about your success, and you want to cut parts cost, don't overlook scrap TVs, radios, surplus computer boards, etc. Usually nothing more than an ohmmeter check is

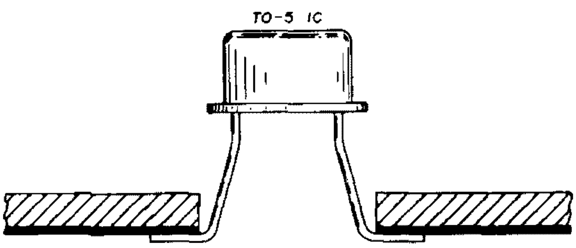


fig. 10. This method of mounting TO-5 packaged ICs allows them to be removed from the circuit board without damage.

necessary. With this type of construction you can use components with very short leads. The receiver in fig. 1 was built with almost all scrounged parts and represents a total cost of about $7.00.

soldering

Your soldering techniques are very important for good PC board work. The economical Ungar soldering iron with a 37- to 47-watt heating element and 1/8-inch (3-mm) wedge-shaped tip is a good combination. An SCR controller or Variac can be used to regulate the heat. For smaller foil areas you can cut the voltage back to about 70 volts, but for large foil areas the iron will require full voltage. Use small diameter solder. Either 0.032-inch or 0.040-inch (about 1 mm) is satisfactory.

Clean the leads of the components before mounting them through the board holes and place the tip of the soldering iron against both the board and lead as shown in fig. 12. Feed a little solder to the same junction to set up a thermal path from the soldering iron to both lead and foil. Then wipe the solder in a circle around the junction. The solder should creep out evenly all around and form a nice fillet as shown in fig. 13A. If you get a lumpy joint as shown in fig. 13B it could be due to lack of heat, dirty board or lead, dirty iron, excessive solder, or a combination of these items. It may or may not be a good joint.

On large copper areas the heat is dissipated so fast that it might require a heavier iron. As soon as the joint is made, wipe the tip of the iron on a damp rag or sponge. Clean the soldered joint with isopropyl alcohol (the discount drugstore variety is ok). The flux residue will absorb moisture and form flux bridges between adjacent foil strips. The alcohol will leave a whitish film which can be removed later by washing the completed board with plain soap and water.

For cleaning the boards I use the inexpensive acid brushes obtainable from good hardware stores. I cut the bristle length in half to stiffen the brush. If your hardware store doesn't stock these brushes, try your local sheetmetal shop. They are a common tinsmith's tool for brushing flux onto a joint.

The foregoing instructions may sound tedious and unnecessary, but you will soon find that they do not take any great deal of time, and the results are worth the effort. A dirty soldering iron and a dirty lead will require many seconds of heating to get a satisfactory joint. Often the first impulse is to feed in more solder. The end result is, commonly, solder so overheated it crystallizes into a grayish mass, may

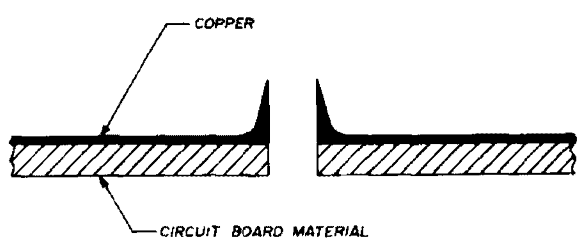


fig. 11. When drilling printed-circuit boards, a dull bit will produce an excessive burr as shown here.

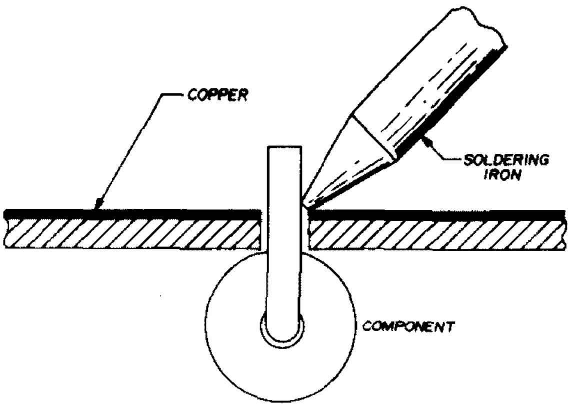


fig. 12. When soldering components to a printed-circuit board place the tip of the soldering iron between the component lead and the copper foil.

not be bonded to the lead, and a copper foil strip so overheated it peels off the board. Follow the four main rules and save trouble:

1. Absolutely clean board, no fingerprints.
2. Keep the soldering-iron tip clean.
3. Use the right heat to get a nice fillet on your soldered joints with little solder buildup.
4. Clean the finished joint.

Incidentally, too little *or* too much heat will produce a grayish, porous solder joint. A proper joint is shiny. If you want to clean up a lumpy joint, you can wick

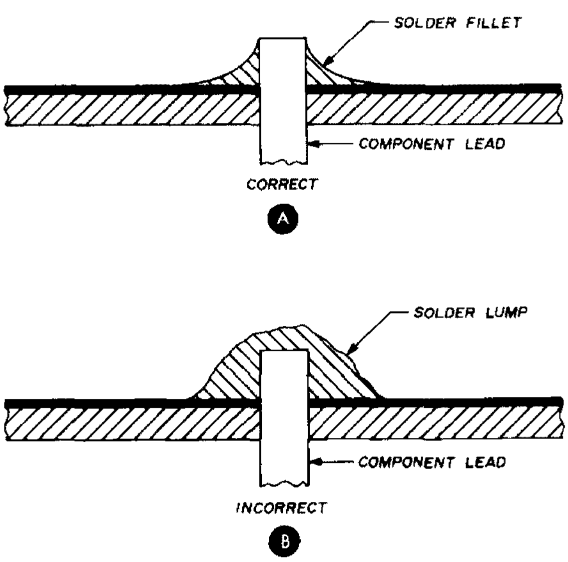


fig. 13. A properly soldered component lead will produce a smooth, shiny fillet around the lead as shown in (A). A lumpy, gray blob as in (B) is usually a poor electrical joint.

the solder nicely as follows: Mix a solution of rosin and alcohol. Dip the end of a small diameter shield braid (about 1/8 inch or 3 mm) into the flux. Lay this fluxed braid on the joint and place a clean, hot soldering iron on top of the braid – the excess solder is sucked up into the braid. This solder wick is obtainable in rolls of untinned braid saturated in flux, but you can find lengths of suitable braid in shielded leads to phonograph pickups, on volume controls of some TV sets, etc.

Do not try to use the so-called noncorrosive soldering pastes! Their flux residues are difficult to get rid of. A tiny quantity in a lead hole can become an insidious time bomb.

Those of you who started in radio along with me – way back when – re-

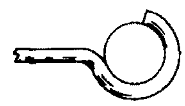


fig. 14. Proper method of wrapping a component lead around a terminal.

member the old cardinal rule was: First make a rigid mechanical connection and then solder a joint. Leads were tightly wrapped round and *round* a terminal before soldering. Many fellows applied plenty of solder for extra strength – particularly men who worked in the marine or mobile services. Then came the space and missile age, and the need for extreme reliability in soldered connections. The extreme vibration and G forces soon proved the old ways were inadequate. For high-reliability work you are forbidden to wrap a wire completely around a terminal. You do it as shown in **fig. 14** and use a gimmick like an alligator clip on a rubber band to hold the wire secure against movement until the solder cools. You use only enough solder to form a fillet, and the lay of the strands of wire show clearly above the fillet.

Of greatest importance is the appearance of the solder. It must be shiny, indicating correct heat and no movement of the wire while cooling. This insures maximum strength in the joint, and low-

est resistance. With excess solder, an inspector can't be sure of the quality of bond to the strands of wire, so you wick it, clean it and do it over again. I don't mean to imply that you should produce aerospace quality boards, I merely want to stress the cleanliness requirements. By following those four little steps you will find it easy to get the one feature you *do* want: Shiny solder joints. Solid-state devices are relatively high-current, low-voltage devices. A poor solder joint is more likely to be troublesome under those conditions.

And don't give up the ship because your eyes aren't what they used to be. I just finished a one year stint on an experimental spacecraft, and the stalwarts on that job were all in their fifties, members of the bifocal crowd, and I had been away from Hi-Rel work for many years.

Now for a few tips: Install your semiconductors *last*. That way they are heated only once. Don't be afraid to leave full-length leads on the transistors on the universal board. It won't degrade performance, and until you are used to PC board work it will help you avoid overheating transistors.

If your station is close to the seacoast you may want to protect your boards from salt-air corrosion. Mask the edges of the board and spray the foil side with clear acrylic.

Finally, use a regulated power supply with low ripple voltage. If you have a junkbox full of surplus semiconductors, the power supply in fig. **15** will be quite adequate. Basically it is a pre-regulator circuit feeding a simple capacitance-multiplier circuit. The result is a fair degree of regulation and very low ripple. A supplementary benefit is that the pre-regulator will provide considerable immunity from power-line transients and voltage variations.

For checking vfo drift you need a supply which is stable with respect to power-line variations. This circuit will provide 12 volts at about 300 mA, with some voltage droop at loads over 300 mA. The circuit can be simplified by eliminating the pre-regulator (C2, R1, CR1) and changing R2 to 470 ohms, 2 watts. Since the transistor will be dissipating about 3.5 watts it can be mounted on the chassis to provide a heatsink. The zener, CR1, will dissipate about 6 watts under no-load conditions so it should also be mounted on a heatsink capable of dissipating this kind of power. Scrap car radios often can be a source of such heatsinks. If your junkbox doesn't contain the transistors and zener diodes, it may be more economical, simpler and less bulky to purchase a three-terminal voltage regulator for 12 volts.

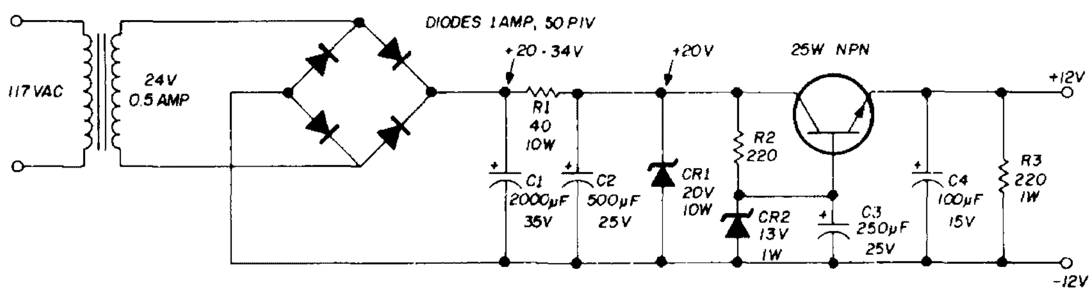
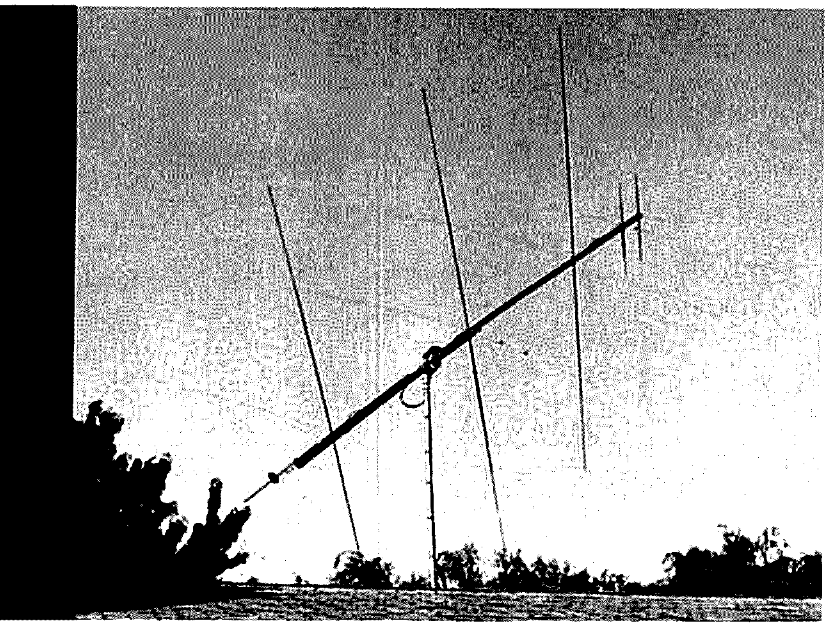


fig. 15. Regulated power supply with low ripple voltage suitable for many solid-state circuits.

conclusion

Although I still build a fair amount of vacuum-tube gear, I find solid state a great deal more fun. I have never been able to obtain such high performance as easily in homebrew gear. Construction today is less a machine-shop job and more of a card table variety of work. I hope these notes will help you, too, to have the pleasure I've found in this field of tinkering.

ham radio



automatic azimuth/elevation antenna control for satellite communications

How to build an automatic az-el antenna control system

George R. Bailey, WA3HLT, Bethesda, Maryland 20014

Anyone who has operated through Oscar 6 using a directional receive antenna is aware of the difficulties involved in spotting the transmitter, conducting a coherent conversation, keeping the log and orienting the antenna. This article shows how to eliminate one of these problems, antenna orientation.

The array used here consists of two 3-element, 10-meter receive antennas and two 2-element, 2-meter transmit antennas, all on the same boom. The 10-meter antenna beamwidth is about 60 degrees so alignment, while not critical, has to be readjusted numerous times on any pass of the satellite. This procedure became tiresome, and finally resulted in the design and construction of the automatic azimuth-elevation rotor control system described here. A block diagram of the system is shown in fig. 1.

This particular rotor control was designed to handle a CDE TR-44 azimuth rotator and an Alliance C-225 elevation rotator. However, any antenna rotators with position-sensing pots can be adapted for use. Basically, the system consists of a cassette tape recorder playing pre-recorded digital information into the rotor control where it is decoded and loaded into two shift registers. The register output is then converted into two analog signals which are propor-

tional to the desired antenna azimuth and elevation. The desired position signal is compared with actual rotor position, and if the error is excessive, the rotators are driven to null this error. Manual operation of either rotator is still possible by use of the mode switch and position metering.

pre-recorded information

Approximately every two minutes, every minute for close passes, a ten-bit binary word is fed to the rotor control. A zero is represented by a 1585-Hz tone, a one by the absence of this tone. A shift signal is also required, and the presence of a 228-Hz tone is used for this purpose. Six bits have been allowed for azimuth information and four bits for elevation. Thus, the minimum change in antenna position is six degrees about either axis. Following each word is a 700-Hz tone which initiates rotator position correction. A typical tape format is given in fig. 2. Bit spacing is not critical, but should be no faster than five times that shown.

At least two external audio oscillators are required to record the tapes. The circuit used to control the audio signal generators is shown in fig. 3. Presently, due to operating schedule limitations of Oscar 6, only South to North passes are available. Consequently, five cassettes, 30 minutes on each side, provide sufficient alignment accuracy for any useable pass from my station.

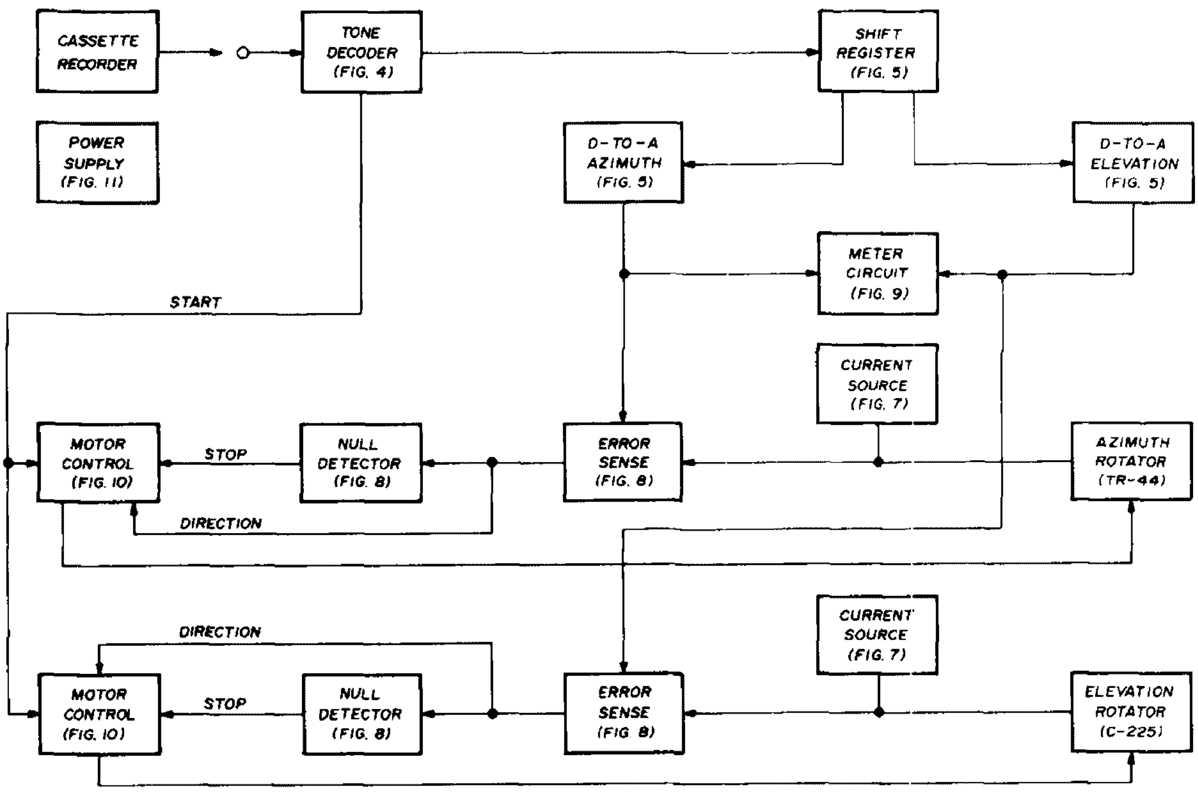


fig. 1. System block diagram of an automatic azimuth/elevation control system designed for satellite communications.

control circuit

The interface between the tape recorder and the TTL logic circuitry is shown in fig. 4. The peak frequency of each selective amplifier is shown on the schematic. Since close-tolerance components are not used, your frequencies

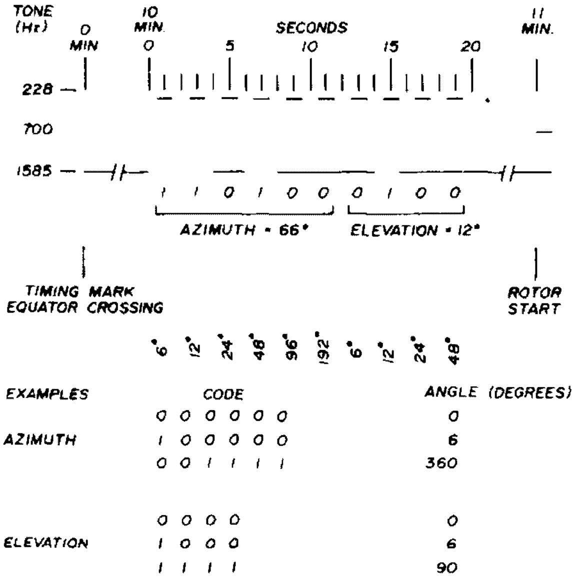


fig. 2. Tape format for cassette tapes used to control the automatic az/el system. A zero is represented by a 1585-Hz tone, a one by absence of this tone. A 228-Hz tone is used to shift the digital information from the shift registers into the D-to-A converters, and a 700-Hz tone initiates rotator position correction. A tone encoder for encoding the cassette tapes is shown in fig. 3.

may be somewhat different. As a result, it is best to build and test at least this part of the rotor control before recording the program tapes. There is one half of a 558 op amp left over which, if you wish, can be used as an audio input buffer. All three of the amplifier-drivers produce a negative-going pulse when the appropriate tone is applied.

The shift registers and digital-to-analog converters are shown in fig. 5. The unmarked resistors must be hand picked to insure not their absolute value, but their relative value. The condition to be met for the elevation D-to-A converter is that: R2 = 2R1, R3 = 4R1 and R4 = 8R1 where R1 is approximately 4700 ohms. Similarly, for the azimuth D-to-A converter: R6 = 2R5, R7 = 4R5, R8 = 8R5, R9 = 16R5 and R10 = 32R5 where R5 is approximately 3300 ohms. The selection of parts to meet these conditions is done with the aid of the test circuit shown in fig. 6. The 7496 ICs shown in fig. 5 are unplugged, and the leads indicated in the test circuit are patched to the 7496 sockets. Next turn both ramp minimum pots to zero, and the ramp size pots to mid-range. An oscilloscope is then connected to first one and then the other desired position outputs. The observed stepped ramp is adjusted for linearity by changing resistors R1 through R4 for

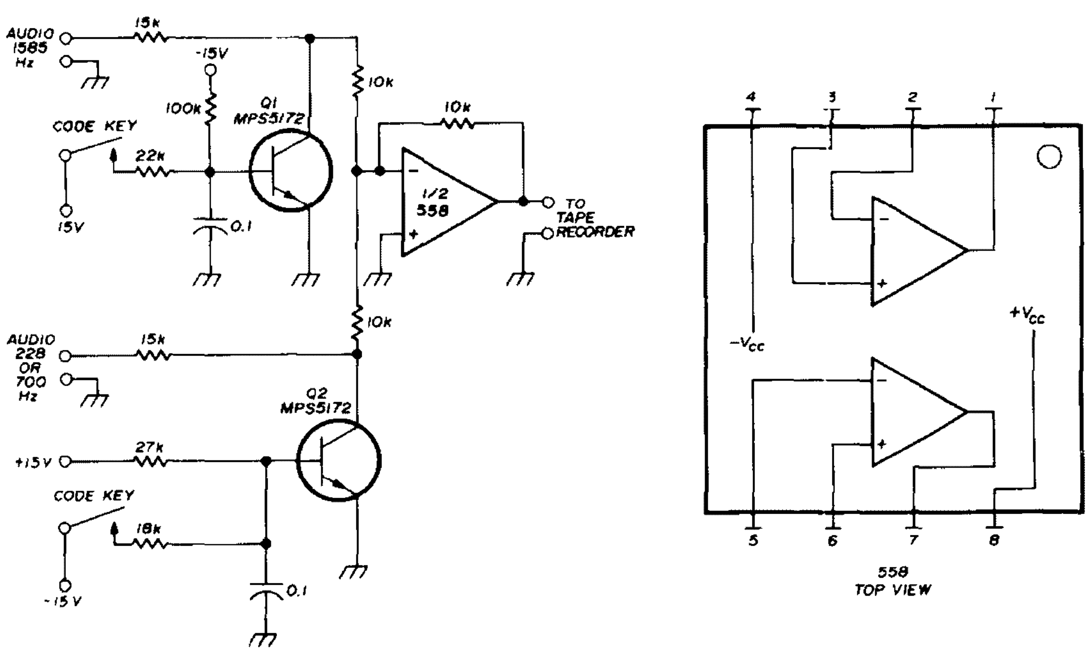


fig. 3. Tone encoder for encoding the cassette tapes required for the automatic az/el system. Tape format is shown in fig. 2.

the elevation ramp and R5 through R10 for the azimuth ramp. After linearity is established, the ramp size pots are adjusted for an elevation ramp size of 1.25 volt peak-to-peak, and for an azimuth ramp size of 5.25 volts peak-to-peak. The ramp minimum pots will be adjusted later.

Two identical current sources are used in the system, one for each rotator resistor and adjust the current pot for 10 mA output current. Open the short across the load resistor and adjust the linearity pot again for 10 mA. If the milliammeter has significant internal resistance, it may be necessary to repeat the above steps.

The null detector circuit is shown in fig. 8. The output of U9 is an indication of the sense of the error between

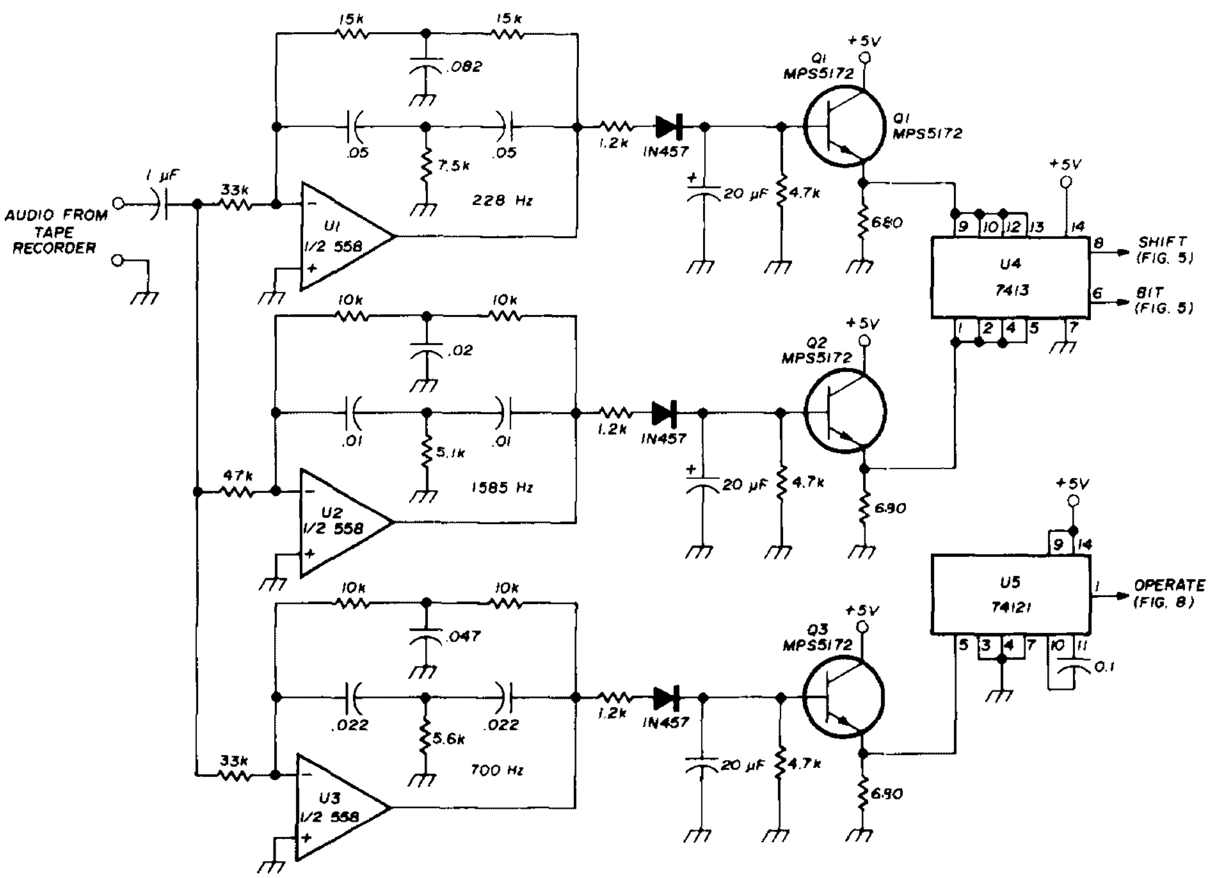


fig. 4. Tone decoder and TTL driver interfaces the cassette tape recorder with the shift registers and D-to-A converters.

position sensing pot. Exactly 10 mA is passed through each 500-ohm pot, regardless of its setting. Thus the return voltage from the sensing pot in each rotator is proportional to that rotator's position. The current source schematic is shown in fig. 7.

To adjust the current source, temporarily connect a 560-ohm resistor in series with a milliammeter from the output to ground. Then short the desired and actual rotator position. Op amp U10 drives the rotator direction relay so that the next time the rotators start, the error is reduced. As the antenna turns, the error decreases and eventually passes through zero. This zero crossing is converted by a Schmitt trigger, U12, into a fast risetime edge which is used to turn off the RS flip-flop U11C - U11D. An identical null detector is used for the second axis

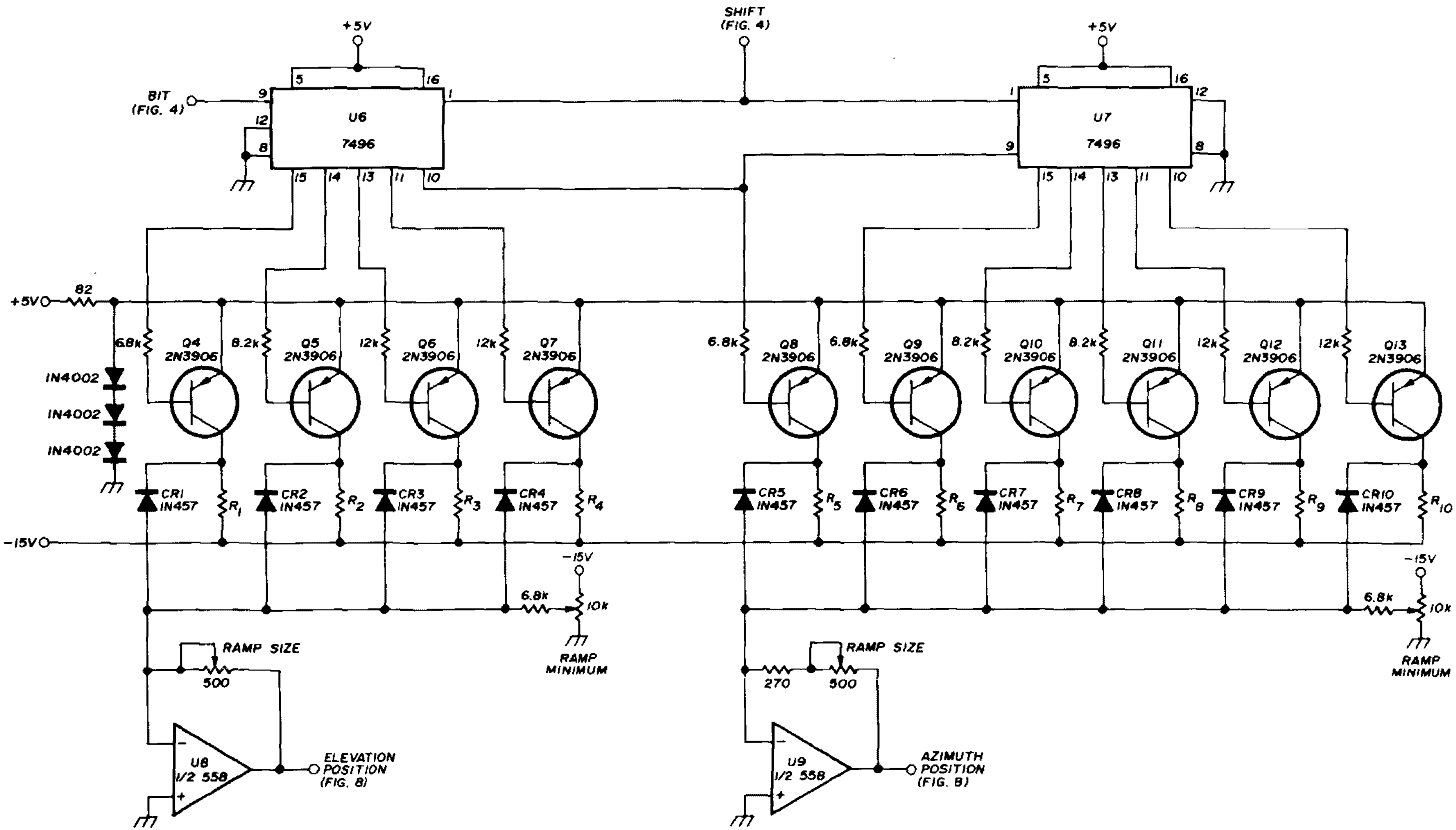


fig. 5. Shift registers (U6 and U7) and digital-to-analog (D-to-A) converters. Transistors Q4 through Q13 are all 2N3906s; diodes CR1 through CR10 are 1N457s. A simple test circuit for aligning the D-to-A converter is shown in fig. 6.

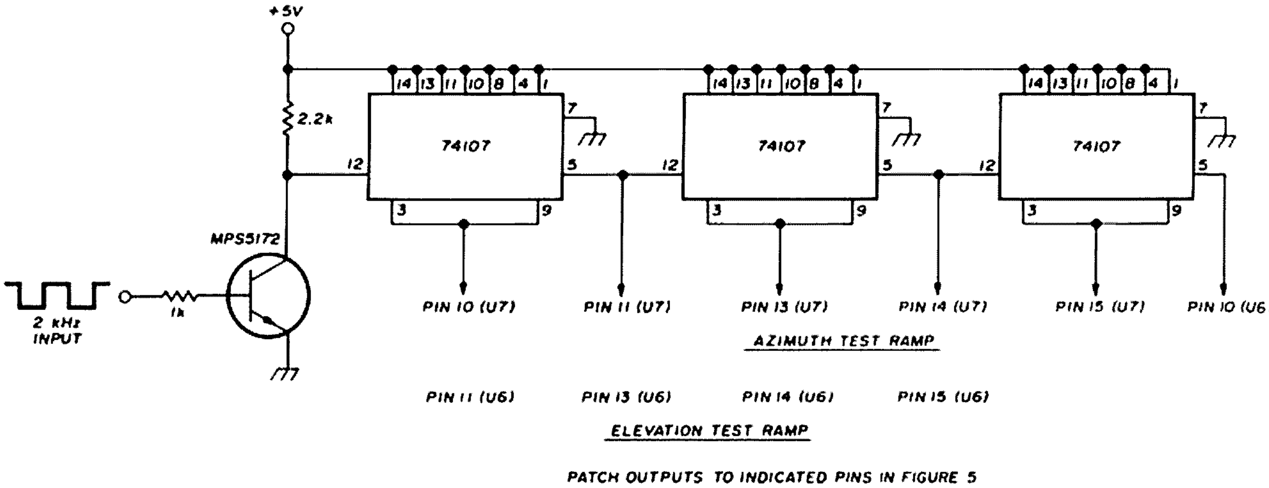


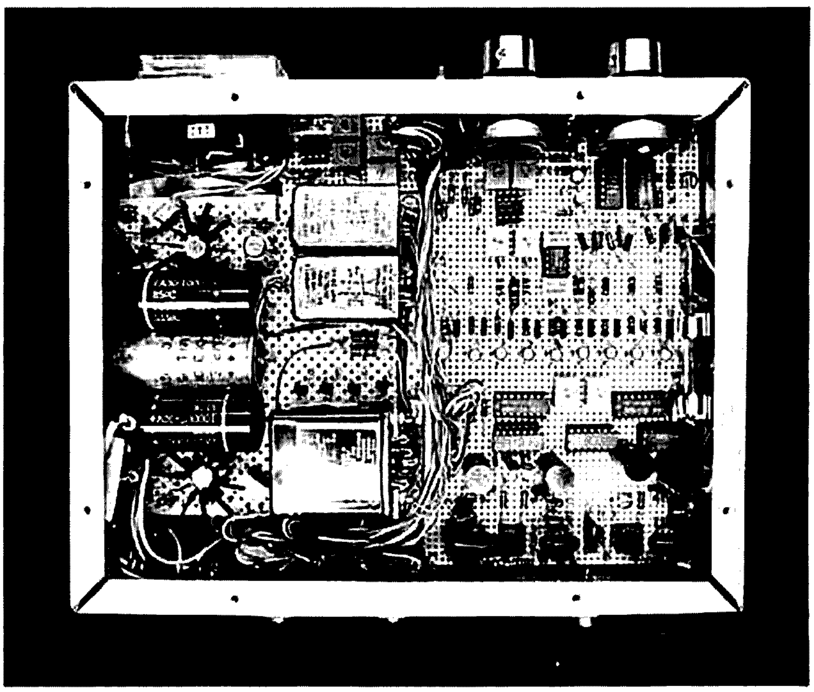
fig. 6. Test circuit for aligning the D-to-A converter of fig. 5. Outputs from test circuit are patched to indicated IC pins in fig. 5 (see text).

which uses the remaining half of U12. The RS flip-flop of each null detector controls the appropriate motor-run relay so that when a null is detected, the rotation about the axis indicating zero error stops. Since the rotator motors are turned off at the error null, the antenna slightly overshoots its desired position.

When the mode switch is in the manual position, the D-to-A converter output is ignored and an externally adjustable pot is substituted, allowing manual entry of the desired rotator position. The rotators are driven to null when the appropriate manual start button is depressed.

The metering circuit is shown in fig. 9. In the azimuth position the meter reads 0 to 400 degrees, and in the elevation position the meter reads 0 to 100 degrees. Calibration will depend on the specific rotators used, and will be described later.

The circuit in fig. 10 is for use with the CDE and Alliance rotators. Other rotators may require circuit modifica-



Layout of the automatic azimuth/elevation control circuitry, showing location of major components.

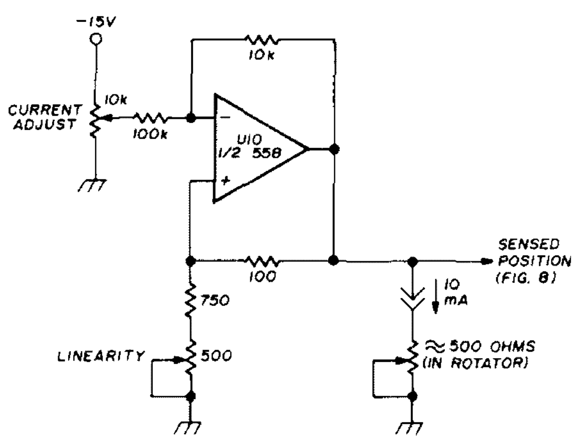


fig. 7. Current source. Two are required, one for azimuth, the other for elevation.

tions. The relays were obtained surplus, and have 28-volt coils with a pair of 5-amp contacts.

One modification is needed on the TR-44 rotator. If you have a series-1 or -2 rotator, remove the sensing pot lead attached to terminal 7 and move it to the previously unused terminal 8. If you have a series-3 rotator, remove the motor common lead from ground and connect it to the previously unused terminal 2. For both axes of rotation, when the rotator direction relay is energized the rotator should rotate in the direction that increases the return voltage from the position sensing pot. If this is not the case, reverse the motor winding leads at the control unit.

power supply

Three separate regulated supplies are used in the system. The +15 and -15 volt supplies use 723 regulator ICs with a finned heatsink on each. The +5 volt supply uses another 723 IC in conjunction with a 40312 pass transistor which is screwed to the chassis. The circuit is shown in **fig. 3**. After the power supply section has been built and adjusted, it is important that the +15 and -15 volt regulators not be readjusted after completion of the remaining circuits. If these voltages are changed, recalibration of the entire system will be necessary.

calibration

Connect both rotators to the rotor

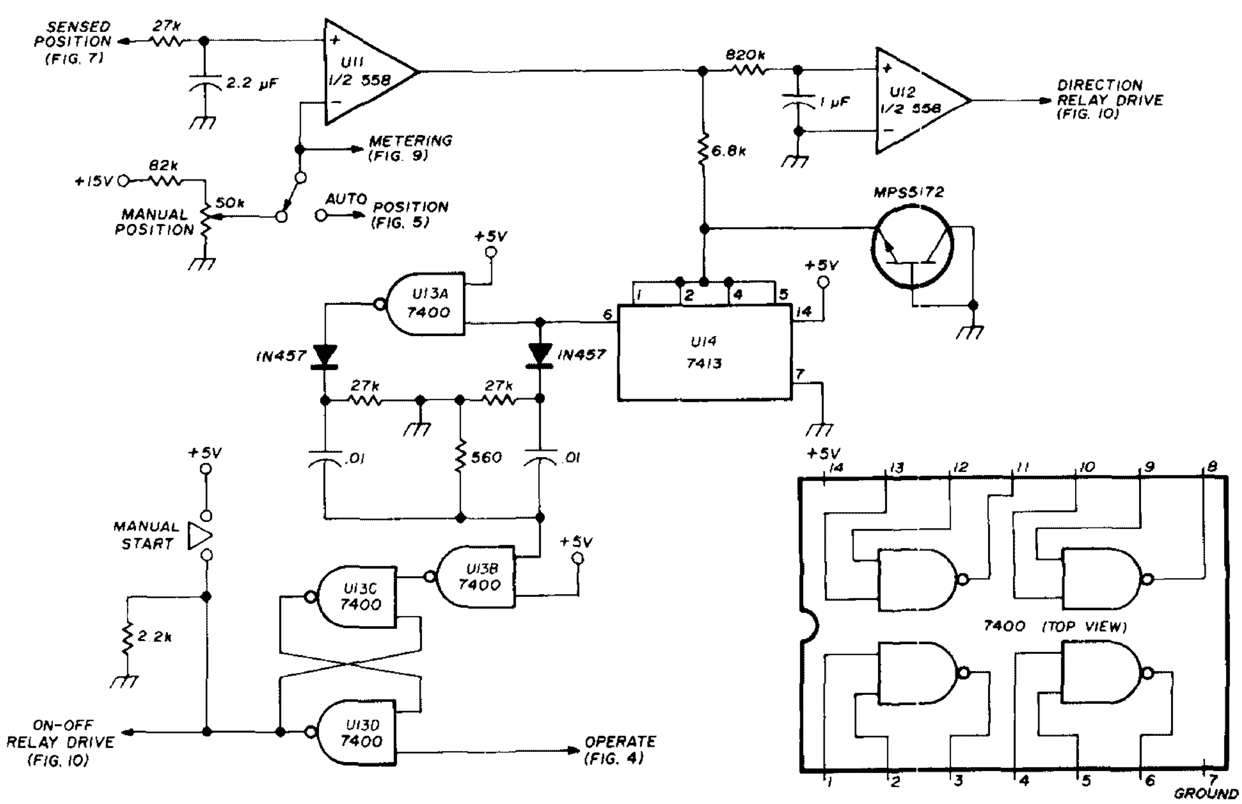


fig. 8. Null detector. Two circuits are required, one for azimuth, the other for elevation.

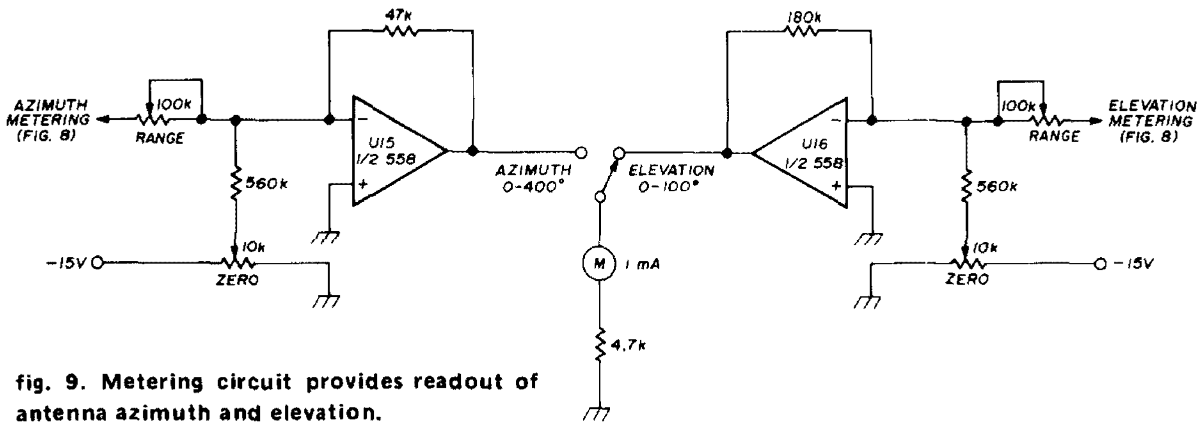


fig. 9. Metering circuit provides readout of antenna azimuth and elevation.

control. Set the mode switch to automatic. With a test tape, enter all zeros into the shift registers. Push the azimuth rotor start button and allow the rotator to stall. Now, slowly rotate the azimuth ramp minimum pot until the azimuth run LED extinguishes. Next, enter 0011111111 into the shift registers and allow the azimuth rotator to either finish rotation or stall. If the rotator is stalled, retouch the azimuth current source adjustment until the azimuth run LED extinguishes. If the rotator stops before rotating 360 degrees, decrease the azimuth current source current and restart the rotator. Continue doing this until the rotator stalls. Then increase the current slowly until the azimuth LED turns off. Again enter all zeros, and then adjust the azimuth meter zero pot for zero meter reading. Again enter 0011111111, and adjust the azimuth meter range pot for a meter reading of 360 degrees (400 degrees full scale).

The rotor control will position the elevation rotator over a 90-degree segment with zero degrees displaced from the low sensing pot resistance end of

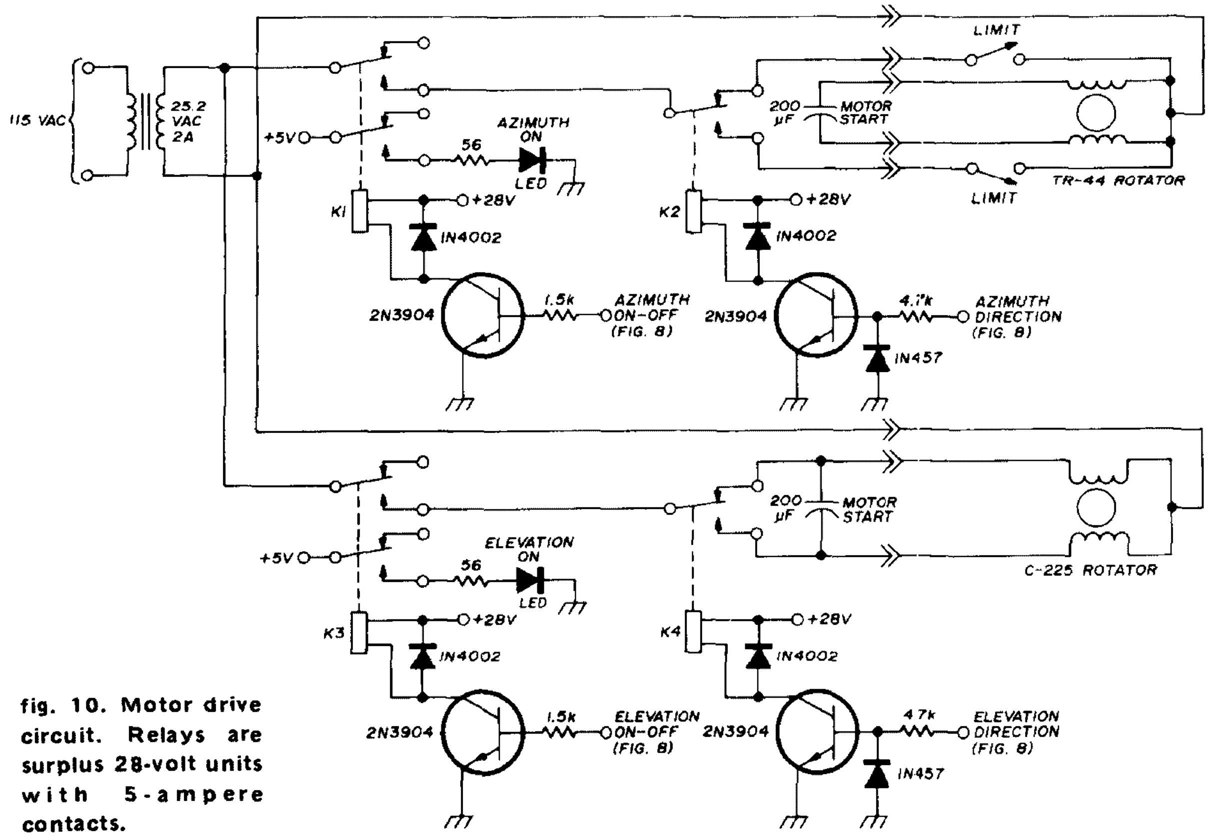


fig. 10. Motor drive circuit. Relays are surplus 28-volt units with 5-ampere contacts.

rotation by as much as 30 degrees. Mark the desired zero-degree point on the rotor shaft, somewhere in the first 30 degrees of shaft rotation. Mark another point on the shift corresponding to the 90 degree position.

Enter all zeros in the registers and initiate the elevation rotator. The rotator should stall. Turn the elevation ramp zero pot through the point that the elevation LED extinguishes. Restart the rotor. It should stop somewhere between the limit of rotation and the desired zero-degree mark. Continue readjusting the elevation ramp zero pot, and restarting the rotator until it stops at the desired zero-degree mark.

Now enter 0011111111 in the registers and initiate the elevation rotor. It should stop near the desired 90-degree position. Readjust the elevation current source current until the rotor does stop at the desired 90-degree point. Again enter all zeros and adjust the elevation meter zero for zero meter reading. Finally, enter 0011111111 and adjust the elevation meter range pot for a 90-degree meter reading (100 degrees full scale).

operation

For manual operation the mode switch is set to *manual and the azimuth* and elevation pots are adjusted to the desired antenna position. Start the rotators with the pushbuttons. Rotation is complete when the LEDs extinguish. For automatic operation simply place the mode switch in automatic and select the pre-programmed tape most suitable for the next satellite pass. When the satellite crosses the equator, start the tape at the zero-minute timing mark. Then forget about the antenna and enjoy good solid communications.

ham radio

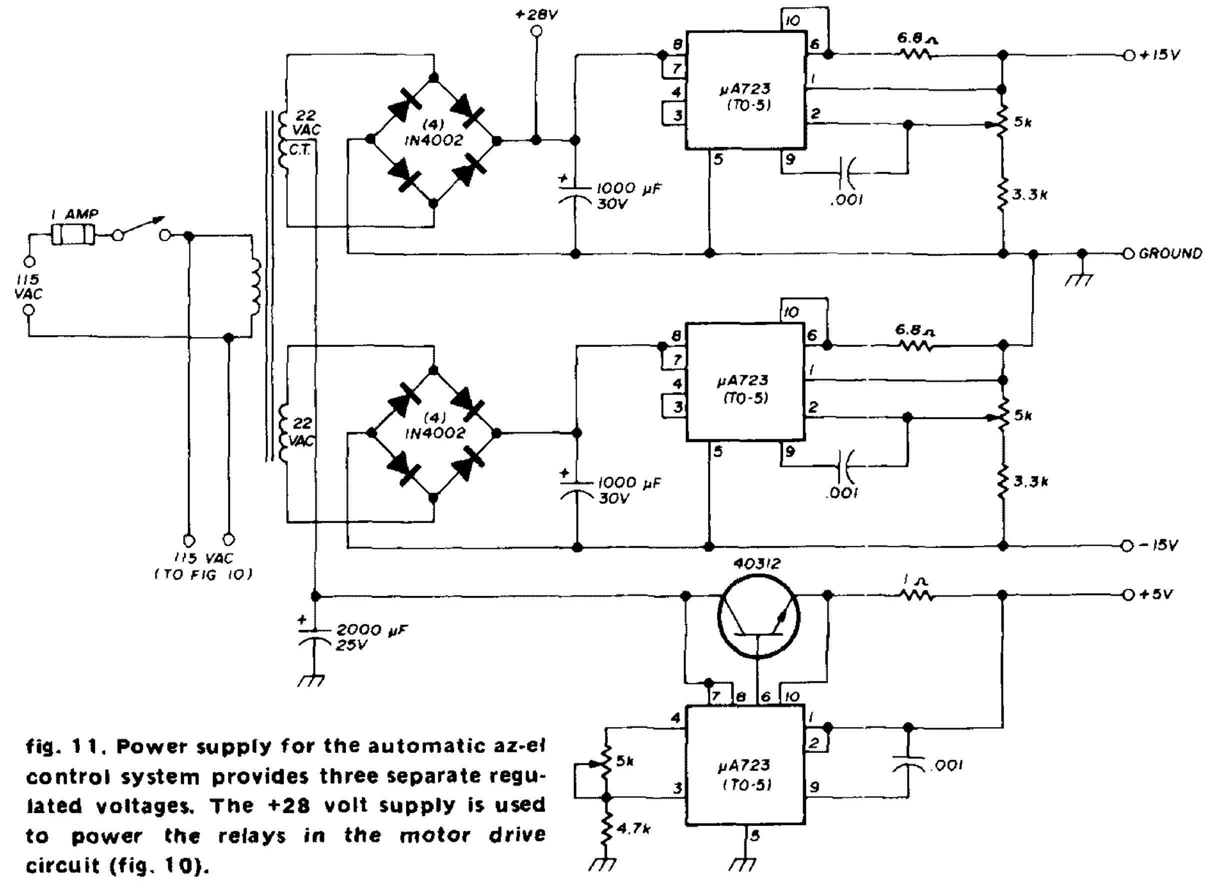


fig. 11. Power supply for the automatic az-el control system provides three separate regulated voltages. The +28 volt supply is used to power the relays in the motor drive circuit (fig. 10).

audio oscillator

Wide range sine, triangle and square-wave generator using the versatile Signetics NE566 integrated circuit

Jim Ashe, W1EZT

If dollars are scarce and shrinking on your workbench but you need a utility audio oscillator, find a Signetics NE566 function-generator IC. Then put it into a simple circuit to get triangle, approximate sine, and square wave outputs over 10 Hz to 25 kHz, in two bands. That basic range extends to 200 or to 500 kHz, if you add one component to the circuit described here.

Total cost for your audio oscillator could run as low as five dollars, if you're an efficient shopper or your junk box is well stocked. Its performance won't be critically sensitive to parts quality – the original model proves that.

The NE566 is a versatile IC, almost as easy to find as the μA741, if it's priced higher yet. Check the electronics magazines to find several sources at under three dollars. When you're buying, the DIP package is best. For NE566 details, see **fig. 1**.

Its 1-MHz maximum is well above what's needed on most amateur workbenches. The band is set by one capacitor at pin 7. Pin 6 offers a 25:1 control range, depending upon resistance, and pin 5 a 2:1 range, depending upon voltage. The circuit described here uses both together for the 50:1 range needed to cover 10 Hz to 25 kHz in two ranges.

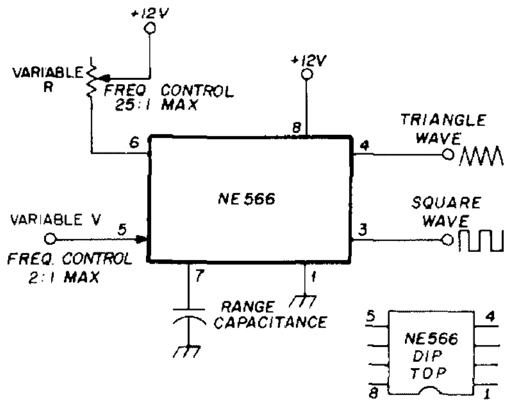
Looking at outputs, the triangle wave is best for all-around application since it's nearly a sine wave. See power spectra facts in **table 1**. There's also a good square wave, if you're interested in digital work.

basic circuit

Fig. 2 shows the basic NE566 oscillator schematic. Its maximum 25-kHz output is 40 times under what the NE566 *can* do, which makes parts values noncritical. However, C2 and C3 should be the best quality available. That is because the NE566 uses constant-current circuits to charge and

table 1. Comparative analysis of sine, triangle and square waves (Fourier power spectra).

	sine	triangle	square
f	100%	84.5%	51.0%
2f	—	—	—
3f	—	9.4%	19.0%
4f	—	—	—
5f	—	3.4%	11.0%
6f	—	—	—
7f	—	1.7%	8.0%
8f	—	—	—
9f	—	1.0%	6.2%



Supply requirements	24 volts maximum, 7 mA typical. 12.4 mA maximum with 12-volt power supply
Input resistance	1 megohm typical (terminal 5)
Triangle output	50 ohms, 0.5% linearity
Square-wave output	50 ohms, 20 ns risetime, 50 ns falltime
Operating frequency	$f_o = \left(\frac{2}{R6C7}\right)\left(\frac{+V - 5}{+V}\right)$ R6 = resistance at terminal 6 C7 = capacitance at terminal 7 +V = supply voltage V5 = voltage at terminal 5

fig. 1. Basic specs of the Signetics NE566 IC function generator.

discharge a capacitor, generating the triangle wave output. If the capacitor's value depends upon voltage across the capacitor (ceramic bypass capacitors) or if it features appreciable leakage depending upon voltage (aluminum electrolytics) then there will be some waveform distortion.

The diodes CR1 and CR2 play a dual role. They clip the triangle wave, and they set the clipping level. Uneven clipping can be improved by trying other diodes, and some diodes will generate a better sine approximation than others. You might want to try a variable resistance at R6 while you're at it, but wave shaping doesn't depend strongly on the value of R6.

For some ideas about simplest construction at least cost, look at the breadboard construction shown in the photograph. This is a great alternative to perforated board assembly.

tuning up

Tuneup proceeds in four steps. First, verify a 50:1 tuning range. For this, put about 0.1 μF at NE566 terminal 7. This range can be adjusted by varying R1, R3 or both. You'll find R3 the most influential.

Next, set *minimum* frequency at 500 Hz by cut-and-try at C3, starting with the 0.077 μF value shown. Then set *maximum* frequency at 500 Hz with C4 added, by adjusting the value of C4. Finally, calibrate the tuning dial in 1, 2, 5 steps as shown in **fig. 3**. This calibration looks coarse, but it's adequate for most bench work. For example, in most radio applications you'll zero in on the desired frequency by ear, while the scale

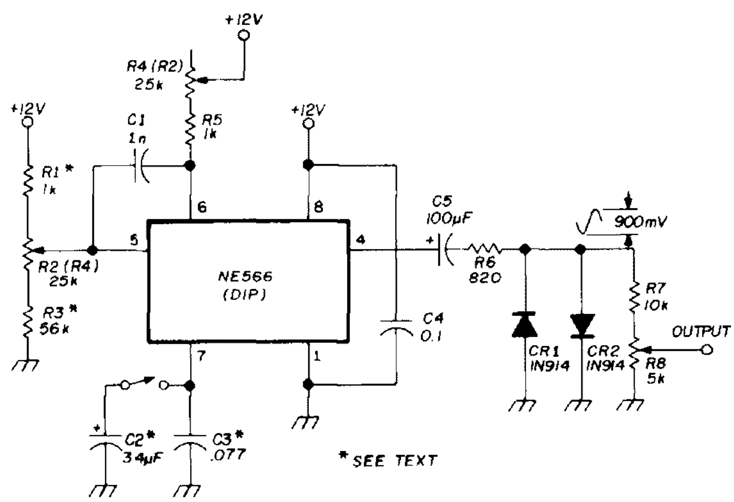


fig. 2. The basic NE566 audio oscillator circuit provides sinusoidal and triangular outputs (see fig. 4) from 500 Hz to 23 kHz. Values of the components marked with an asterisk are set during the tuneup procedure described in the text. Frequency dial for this unit is shown in fig. 3.

provides orientation. For audio you'll be interested in the -3 dB points, which are vague in any case. And for pulse circuit work, perhaps using the NE566's square wave output, your scope displays timing. A large, precise calibration just isn't needed for most work.

The tuning dial shown in **fig. 3** was cut from cardboard, glued to its knob and calibrated in pencil. Then it was taken off the shaft and the calibration inked in with India ink. Another piece of cardboard holds the electrical-tape index. Simple – and good enough!

Fig. 4 shows the original circuit's output. The sine wave is imperfect since two diodes don't provide the best possible wave shaping. The slight difference between rising and falling sides comes from the worst-choice capacitors put in to observe circuit sensitivity to such parts (C2 consists of bypass ceramics, and most of C3 is an aluminum electrolytic). Use better parts, and expect a better signal.

other circuits

Of course you'll have your own ideas about building this circuit. Here are some suggestions you might want to throw in. If the triangle wave is 84.5% fundamental, why not omit the shaping network? That would achieve more output and lower output resistance. See **fig. 5**. Resistor R2 can be 100 to 500 ohms;

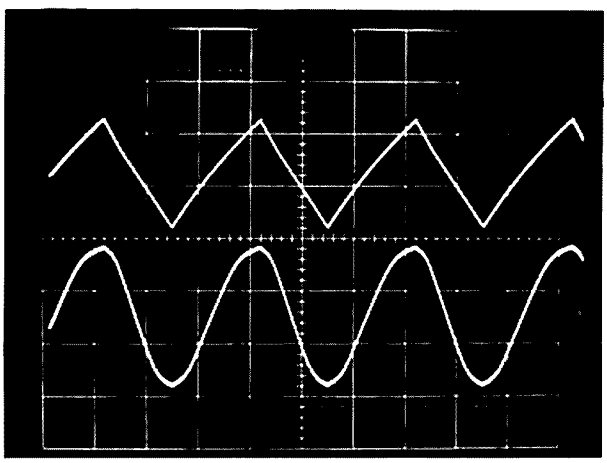


fig. 4. Triangle and approximate sine wave outputs from the NE566 signal generator.

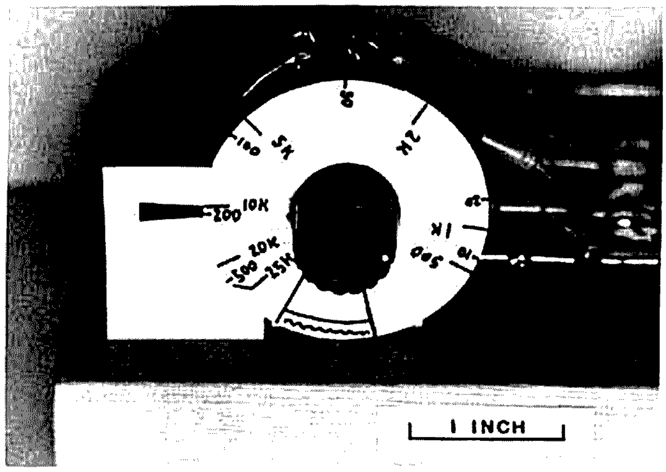


fig. 3. The frequency calibration scale. Detail isn't needed.

R1 as required for amplitude. Total load on the NE566 terminal 4 should stay over a few hundred ohms. At the lowest operating frequency the reactance of C1 is under 10% of the circuit resistance. A good ballpark value is 1000 μF.

An op-amp voltage follower *after* the level control will put output resistance under one ohm. See **fig. 5** again. Since the NE566 triangle wave peaks don't reach the power supply or return levels the μA741 doesn't need a bipolar supply to follow the signal. A large capacitor holds one end of the amplitude pot at signal ground potential and provides a return terminal for the wave forming diodes.

Since the square-wave output is there and many of us like square waves, let's look at that. See **fig. 6**. The square wave is taken out much as suggested earlier for the triangle wave, or a μA741 can be added to reduce output resistance.

The two band-setting capacitors, C2 and C3, could be increased to three. Then a capacitor about 20 times smaller than C3 would give a range of 10 kHz to 500 kHz; the other two ranges might remain the same.

application notes

There are lots of facts around concerning basic audio test work. Here, let's review a couple of the finer points.

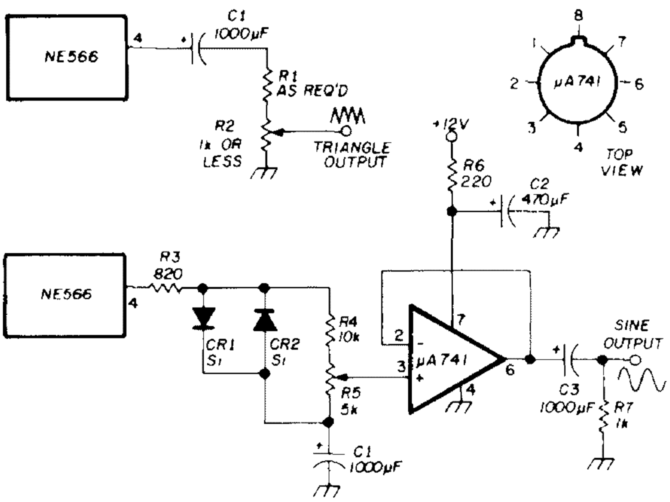


fig. 5. Optional circuits for triangle wave or voltage follower outputs.

Anybody can make a small signal by putting a large resistor in series with a small one, as a voltage divider. However, practical application brings in practical details. Test leads can pick up interference, and this becomes bad in the ac electric field from typically open house wiring. And if a series resistor isn't provided in the voltage divider it's possible to achieve deceptively good signal-to-noise ratios that don't show in real-life application. See **fig. 7**.

Here, long leads from the generator feed a voltage divider situated an inch at most from the small-signal input terminal. Any interference acquired by the leads is reduced along with the signal,

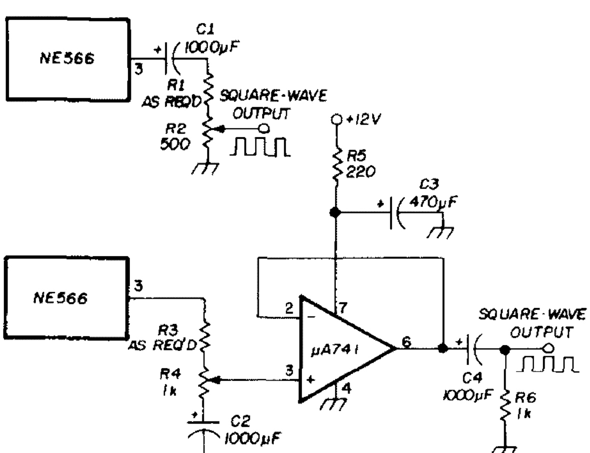


fig. 6. Optional circuits for providing a square-wave output.

keeping the input signal-to-noise ratio high.

Then series resistance R3 imitates the nature of a real signal source, as the driven circuit sees it. Without R3, the small value at R2 loads the tested circuit's input terminals. That shorts out the noise the tested circuit generates in its own components and current flows. Test results taken without R3 or R8 will look very good, but they won't be true results.

power supply

This audio oscillator can be powered from the circuit it's testing, if the

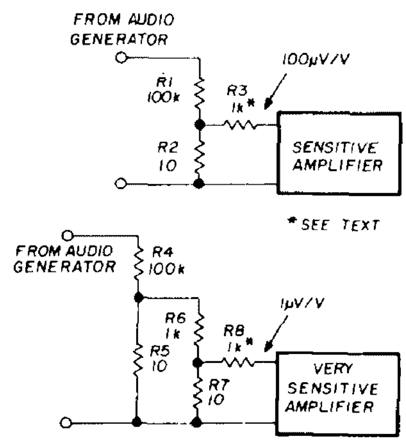


fig. 7. How to obtain interference-free microvolt and nanovolt level outputs.

supply level is 12 volts. That saves finding another power supply — but you must make sure the generator and the driven circuit cannot interact through their common power source.

That's the purpose of R6/C2 in **fig. 5**, and of R5/C3 in **fig. 6**. An operational amplifier can impose a surprising amount of signal on its dc supply line if it's feeding a low-resistance load. These components decouple any roundabout output so that it cannot pass via supply lines through bias networks into a tested circuit, spoiling test results. As an extra precaution, one or two 1000-μF capacitors might be soldered across the generator's supply lines.

ham radio

regulated, variable solid-state high-voltage power supply

Design and construction of a regulated 50- to 300-volt power supply with built-in current limiting and transient protection

Hank Olson, W6GXN, Post Office Box 339, Menlo Park, California 94025

The typical +150 to +300 volt regulated power supply (used for providing bias, screen or plate voltage for vacuum-tube circuits) has, until recently, most often been a tube-type design itself. This has been true because tubes are durable and have been inexpensive in the past. However, series-regulator tubes like the 6AS7G, 6080, and 6L6 are now becoming expensive (nearly $10.00 each), and they age all too fast in regular use. The price of a set of tubes for a typical regulator (6L6, 6AU6, and 0A2) is roughly $14.00. This makes some of the solid-state devices more attractive than they may have previously appeared.

The solid-state version of a regulated high-voltage supply may also have the built-in advantage of adjustable current-limiting, instant turn-on, and (barring accidents) virtually unlimited component life. Some users of the solid-state power supply may miss seeing the plates of the rectifier and series-tube glow red when the output terminal is shorted, as occasionally happens when testing circuits, but that is a minor disadvantage and affects only a few.

The regulated high-voltage supply described here is shown schematically in

fig. 1. It is continuously adjustable in output from 50 volts to 300 volts, at load currents up to 100 mA. The cost of the IC, two high-voltage transistors, and seven diodes (in the regulator portion) is as low as $21.85. This is at least comparable to the price of an equivalent vacuum-tube regulator.

Note that there is a small variable auto-transformer in the primary circuit of the high voltage transformer. These variable auto-transformers are most often called "Variacs" by amateurs, but *Variac* is a registered trade name of the General Radio Company and cannot be properly used in describing the particular variable auto-transformer used in this circuit. This variable auto-transformer is mechanically ganged to the dc voltage-control potentiometer connected to pin 8 of U1.

This technique is one that I first observed on an older commercial tube-type regulated supply (a Hewlett-Packard 710A). The intent is to keep the input-to-output voltage difference, which is impressed across the series-regulator transistor (or tube), nearly constant. In the vacuum-tube model this was done to minimize tube dissipation, increase efficiency and make a smaller package possible without blowers.

In the solid-state version described here, the ganged primary variable auto-transformer and dc output voltage control pot have the same function. However, limiting the voltage across the series transistor is also of considerable importance. This is because the V_{ce} limitation of bipolar power transistors is about 300 volts (for transistors in the price range that most hams would consider reasonable).

If this mechanical complexity seems unnecessary, assume the case where you require 50 volts output at 100 mA. The unregulated input to Q1 would be about 350 volts so there would be 300 volts at 100 mA across this transistor. Not only would this cause 35 watts dissipation in Q1, it would also stress the device to its V_{ce} limit. At load currents lower than 100 mA the input voltage would be even higher and the V_{ce} rating would be exceeded even though less power might be dissipated. With the variable auto-transformer in the primary of the high-voltage transformer, T2, the differential voltage across Q1 will never exceed 100 volts and the power dissipation of Q1 will be held to 5 watts maximum.

regulator circuit

The regulator circuit is designed around the Motorola MC1466L (or MC1566L) which is that company's *floating* regulator. This IC was designed with variable lab-type power supplies in mind and is powered by a 25-volt supply which "floats" (i.e., the power supply has no common connection to ground, but delivers its output only to pins 7 and 14 of the MC1466L IC). The only dc paths to ground that exist for the MC1466L are through the dc voltage-control pot, the load, and back through Q1 and Q2 to the unregulated high-voltage source. Note that transistors Q1 and Q2 act as if they were *one* high-beta transistor since they are connected in the Darlington-pair configuration.

The MC1466L/MC1566L regulator IC is one that has received some adverse publicity, but the past problems apparently have stemmed from lack of adequate transient protection. For instance, it was reported in *Electronic News* some time ago that Power-Mate (manufacturer of regulated power supplies) was suing Motorola over the MC1466L/MC1566L because their power supplies were failing in the field. I have not seen any further word of this problem so it is assumed that the two companies got together and solved the problem.

Next, at work I began to get modular power supplies (Lambda LCS-3 series)

back from the field with IC failures. These ICs were apparently MC1466L/MC1566L types, too, judging by their unique pinouts (although they were marked with an in-house number). A request for assistance from the manufacturer yielded not only advice, but that agree on the method and a complete lack of field failures in the modified modular power supplies) that it seemed prudent to go ahead with a regulated high-voltage supply using the MC1466L/MC1566L. The discussion above is not meant to criticize

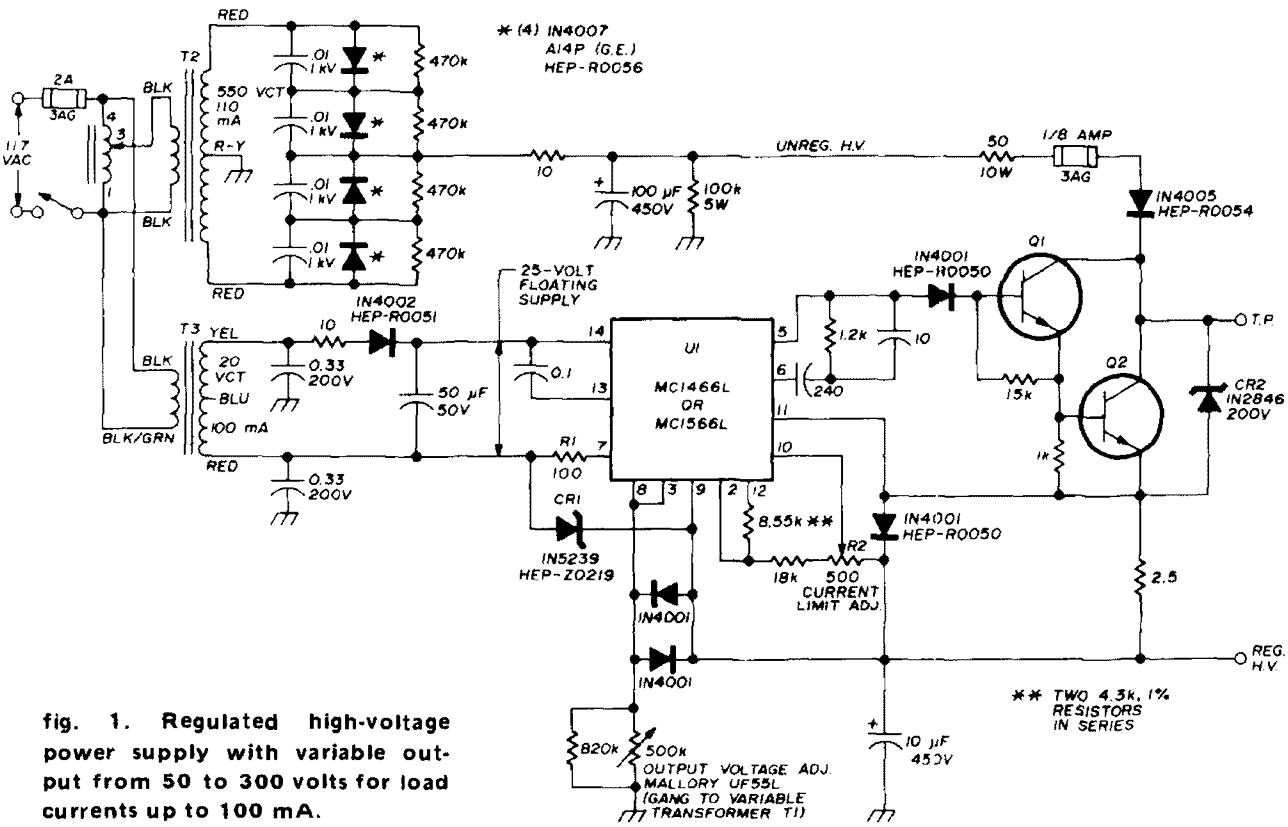


fig. 1. Regulated high-voltage power supply with variable output from 50 to 300 volts for load currents up to 100 mA.

CR2	200-volt, 50-watt zener diode (heatsink to chassis)	T1	0-132 volt, 2.25 A (0.3 VA) variable auto-transformer (Superior Electric 10B)
Q1	Motorola HEP244 or MJE340 (heatsink to chassis)	T2	550 volts center-tapped, 110 mA (Triad R112A or R12A, filament windings not used)
Q2	Motorola HEP707 or MJ413 (heatsink to chassis)	T3	20 volts center-tapped, 100 mA (Triad F90X)

modification kits (free of charge) that solved the problem. Subsequently, in the third edition of the Motorola *Linear IC Handbook* (1973) a note appeared on transient failures. In the fourth edition of this handbook the recommended transient-protective circuit was shown.

The transient protection picture was finally clear enough to me (two sources Motorola, Power-Mate or Lambda, but rather points up how years of field experience are sometimes necessary before all the subtle points of a component are *really* understood. I'm sure that the same was true of the 6L6 and 807 in their early days.

circuit operation

Perhaps a more detailed look at the

circuit in **fig. 1** is in order. As outlined above, transformer T2 derives its primary ac voltage from terminals 1 and 3 of T1, the variable auto-transformer. Since the ac voltage output of all the secondary windings of T2 vary with the setting of T1, the 5- and 6.3-volt secondaries of transformer T2 are not used. The low-voltage transformer, T3, for the floating supply, operates directly from the constant ac line voltage.

The high-voltage rectifier uses four 1000-volt PIV silicon diodes in a full-wave, center-tapped circuit. Since the variable auto-transformer can push the primary ac voltage of T2 to 135 Vac, the ac output can go as high as 320 volts. Thus, the conservative thing to do is to use two 1000-volt PIV diodes in each leg of the rectifier. The standard 0.01 μF disc capacitor and 470k resistor are connected across each rectifier diode to assure equal voltages across each. A 100-μF capacitive input filter is used with a 10-ohm series resistor to limit input current to a safe level for the diodes (and for the capacitor). A 100k bleeder resistor was added across the 100-μF filter capacitor to minimize surprises when doing initial circuit work.

The floating power supply is about as stark a supply as you will find anywhere. It is a simple half-wave type with a silicon diode, 10-ohm surge current limiting resistor and a 50-μF filter capacitor. Since the worst-case current drain from this supply (by the MC1466L) is only 12 mA, not much of a supply is required. The two 0.33-μF capacitors from the secondary leads of T3 are part of the transient protection for the IC.

The 100-ohm resistor, R1, and 1N5239 zener diode, CR1, in the regulator circuit are the primary recommended components to prevent IC breakdown during transients. The 0.33-μF capacitors mentioned above represent *additional* transient protection.

There are some other added protective components that I felt would save the high-voltage transistors Q1 and Q2 in a particular failure mode. These are the 1/8-amp fuse in the unregulated high-voltage lead and a 200-volt zener between the emitter and collector of Q2. To explain this zener and fuse combination, suppose that the supply is set at +120 volts output and current limiting is set at 100 mA. Should the output be shorted, the *entire* unregulated high-supply is impressed across Q2; this would be about 180 volts at 100 mA or 18 watts. The transistor can handle this amount of wattage at least long enough for the operator to switch the supply off.

However, if the supply is set at, say, 280 volts output and 100 mA limiting, a short across the output will place about 330 volts across Q2. This, of course, exceeds its V_{ce} rating and will destroy the transistor. The 200-volt zener across Q2 limits the voltage during an output short condition to 200 volts, and the fuse opens at 125 mA to protect the *zener* from over-dissipation.

A high-current zener diode must be used (the zener must be able to handle at least several times the fuse current for a good part of a second). The 1N2846 is a 50-watt, 200-volt zener that is rated at 200 mA continuous current. The power supply will allow the output to go into current-limiting, providing the output voltage is below about 150 volts. At output voltages between about 150 volts and 300 volts the supply is still protected from a short circuit but a high-voltage fuse will have to be replaced after the short is removed.

construction

The supply is built into an LMB type W-1C box-chassis. This relatively small cabinet has more than enough room for everything; I can't even imagine trying to cram a 5T4, 6L6, 6SJ7 and 0A3 into such a box as we used to do. The voltage control pot (a Mallory UF55L,

made for coaxial use) is mounted on the front panel above the chassis plate. Directly behind this pot and aligned with it, mounted on an angle bracket, is the Superior Electric model 10B variable auto-transformer. The two are coupled together with a piece of 0.188-inch (5mm) drill rod which fits the ID of the hollow potentiometer shaft. The other end of the 0.188-inch rod is bushed back up to 5/16 inch (8mm) with an unused piece cut from the (long) pot shaft itself (this pot has a larger than 1/4-inch [6.5mm] shaft). The 5/16-inch diameter shafting means that the voltage-control knob and one side of the flex coupling (which is coupled to the auto-transformer) must be drilled out to 5/16 inch. A small hole is drilled in each of the two 5/16-inch hollow shaft pieces (one on the pot and the other on the bushing) so that the respective set screws will lock the 0.188-inch rod solidly to the hollow shafts.

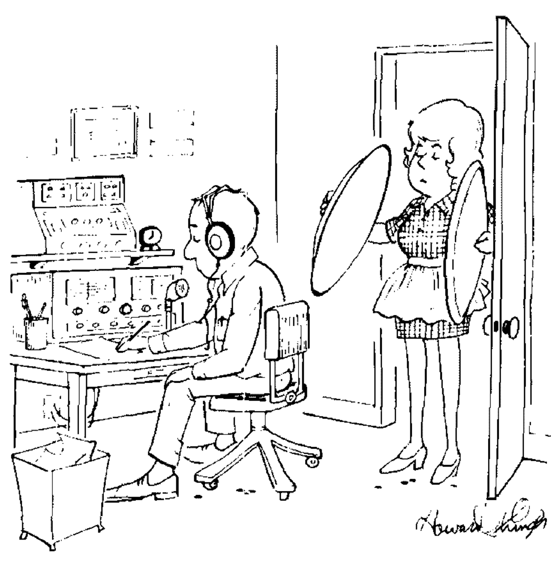
The potentiometer should be set for minimum resistance and the auto-transformer for minimum ac output voltage from terminals 1 and 3 so that turning the output voltage control will increase pot resistance and simultaneously increase the ac output voltage. Then the set screws should be tightened.

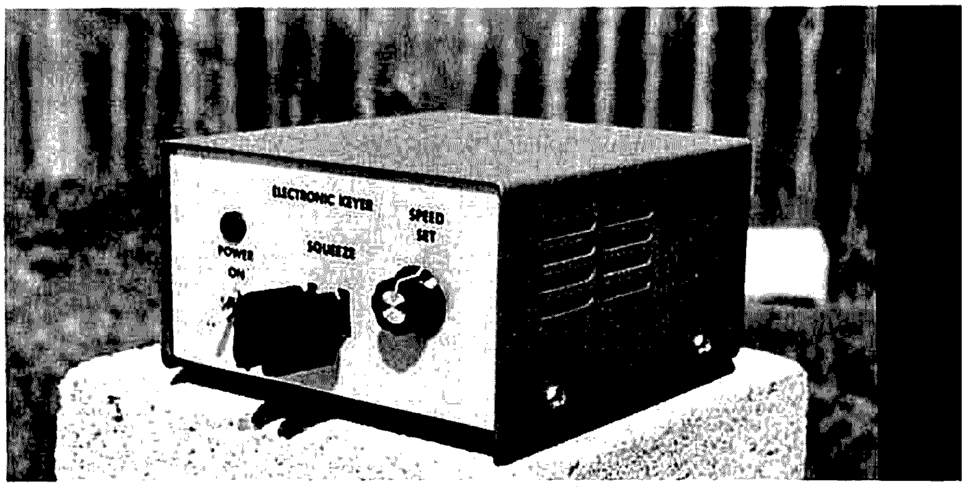
The current limiting control, R2, is placed on the front panel, under the chassis plate near the circuit board with the IC and its associated components. The two transistors and CR2 are mounted on the chassis plate near the rear with mica washers and silicone grease for heat-sinking. CR2 and Q1 are mounted on top of the chassis plate with their pins protruding through this plate, and Q1 is mounted under the chassis. Although these mica heatsink washers are quite thin, they seem to be adequate to insulate the high voltages involved. At normal full-load operation these transistors do not even feel warm to the touch (when touching them, however, remember that they have the *full unregulated high-voltage on them* – turn the supply off and wait until the bleeder reduces the voltage to a safe level).

A couple of other components in the circuit deserve mention. The 8.55k resistor between pins 2 and 12 of the IC is actually two 4.3k, 5%, 1/4-watt resistors in series. If you have an 8.55k, 1% resistor, use it by all means, but 4.3k resistors are easier to find. Also, the 2.5-ohm resistor between pins 9 and 11 of the IC can be a ½-watt, 2.4-ohm or four 10-ohm, ¼-watt resistors in parallel.

There are numerous possible substitutions for the silicon rectifiers used in the supply; just make sure the PIV ratings match those called for. Similarly, the 9.1-volt zener can be most any 8.2- or 9.1-volt type with a 250-mW or larger rating. The high-voltage zener, 1N2846, may be replaced with a 1N3350. The high-voltage transistors Q1 and Q2 can also be replaced by other devices; Motorola has a number of such transistors, mostly in their MJ and MJE series. Other manufacturers (RCA for example) have lines of high-voltage npn transistors which might be suitable.

ham radio





dressing up the siamese paddle

Some packaging ideas for electronic keyers

F.E. Hinkle, WA5KPG, Austin, Texas 78758

Electronic keyers have replaced the semi-automatic bug in many CW stations, and that's a sign of progress. However, a problem common to many home-built keyers is efficient operation of the paddle. This mechanical part of the keyer must provide two switch closures: one for dots and one for dashes. No matter how sophisticated the keyer electronics, the paddle plays an important part in how your CW sounds. An otherwise excellent keyer with a poor mechanical input will cause errors and poor operating habits.

Presented here are some ideas on improving paddle operation and providing a smart, professional appearance to any home-built electronic keyer. The "siamese paddle"[1] is used as an example of what can be done.

The siamese paddle is so-called because two surplus telegraph keys are used to provide the mechanical input to the keying circuit. The keys are mounted back-to-back on a heavy base.

Although this method works fine, the esthetic appearance of the key leaves much to be desired. Why build a good keyer only to have one of its most important features look 30 years behind the times? The refinements shown here will give an overall sharp appearance.

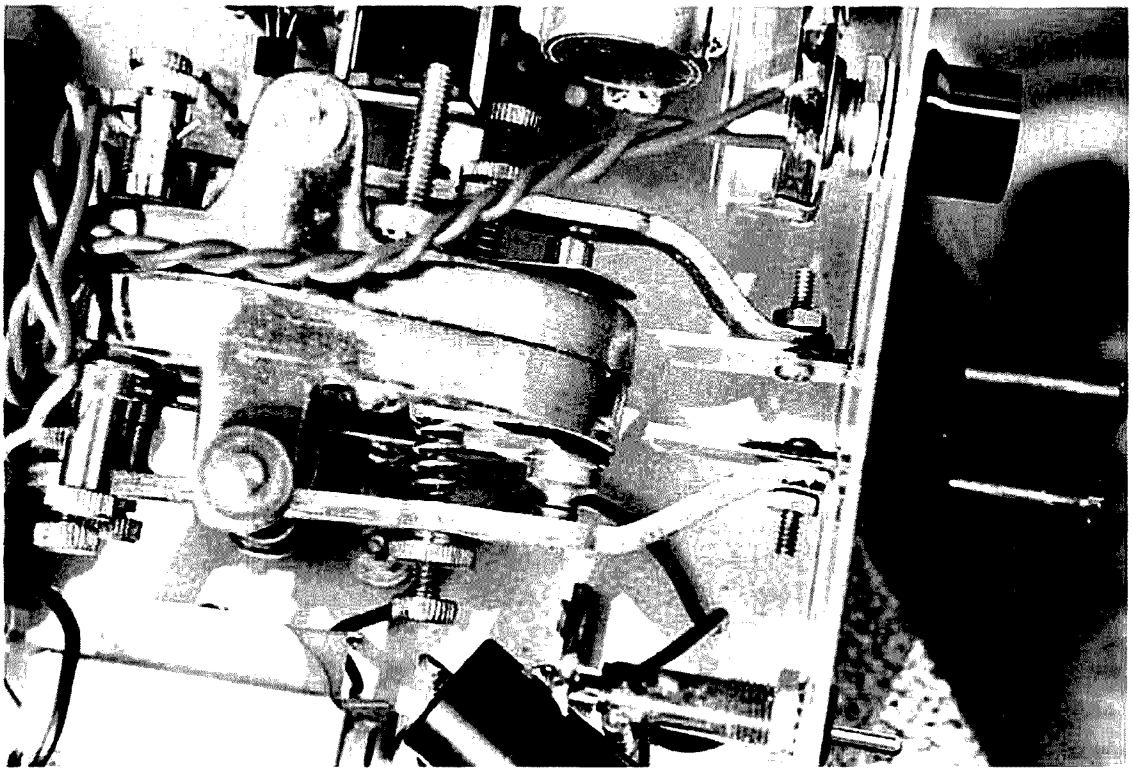
mechanical input

Three types of keys are available on the market today: the surplus telegraph keys (J-38), the ball-bearing pivot Japanese version of the J-38, and a plastic version of the J-38. The surplus and ball-bearing pivot types are better suited to a keyer design because of friction reduction and the multiple adjustments possible. These keys are available for about $3.00, depending on the source. It's important that the keys operate very smoothly and with little effort.

An alternative approach to a telegraph key is a microswitch for the mechanical input element. However, because of lack of adjustments and the extra mechanical considerations this type of key was not used.

packaging

If a siamese paddle could be constructed in an enclosure, the appearance would certainly be enhanced. After finding such a container, it became apparent that it had a lot of wasted volume. A suitable box size for the paddle is 5¼ x 3 x 5-7/8 inches (13.4 x 7.6 x 15 cm).* With the extra space, keyer electronics and power supply fit easily into the container with the keys, making a very neat package.



Two telegraph keys are mounted back-to-back in the center of the cabinet. The ends of the keys are cut off, with fiberglass extension arms added. The handles are slipped over the arms.

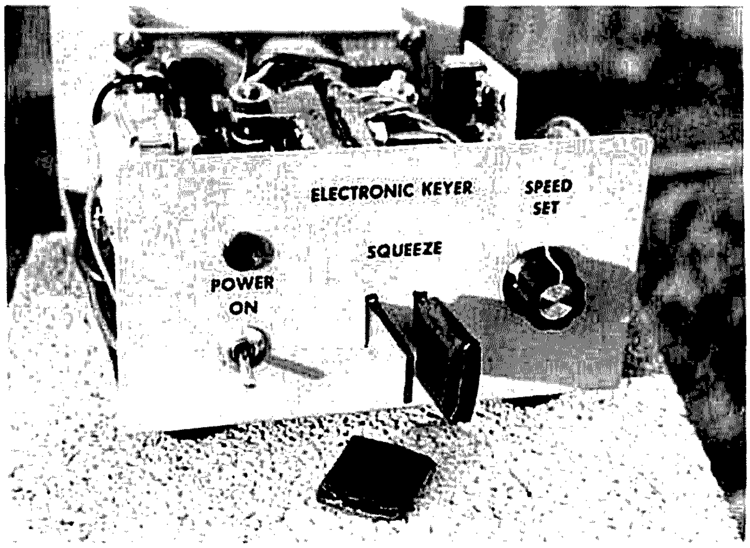
To install the two telegraph keys in the box, a method of mounting must be determined. The easiest approach is to use two bolts and two L-shaped brackets. A 2-inch (51-mm) bolt is used to fasten the keys back-to-back. The two L brackets are then attached to the re-

*A similar box is available from Radio Shack Stores for about $3.40.

maining hole with another bolt of the same size.

The keys may now be mounted at right angles to their normal positions, permitting lateral motion instead of an up-down motion. The keys are quite steady with just two mounting brackets. Since the arms of the keys are longer than necessary, they should be cut off about 1½ inch (38 mm). Two small holes are then drilled into the arms to mount the paddles. Fiberglass printed-circuit boards work well for the paddles. The two arms on the keys should be bent to allow about 9/16-inch (14-mm) clearance between the paddles.

The photo of the keyer interior shows the twin paddle mounted in the middle of the box. Two 1/2 x 1-3/4-inch (13 x 19-mm) fiberglass paddles are brought out 1-1/4 inch (32 mm) from the front plate. The paddle emerges through two small file-slotted holes. Because of the irregular shape of the keys, the power supply is mounted on the left side of the box, while a sidetone circuit and speaker are mounted on the right side. An integrated-circuit Iambic keyer board is mounted with two standoffs directly behind the keys. The back panel includes the output jack, a tune/operate switch, and a sidetone on/off switch. Ventilation holes in the cover allow the sidetone to be heard from the enclosure. Four rubber feet mounted on the bottom of the enclosure hold the keyer steadily during operation.



Front of the paddle with one black handle removed. The handle slips over the protruding arm extension, and the fiberglass arms extend through two small file-slotted holes in the front panel.

finishing touches

Decals were applied to give a final touch to the case. The dressing for the paddles is a plastic slip-on cover to give a better feel. Since the fiberglass paddle is about 1/2-inch (13-mm) wide, a convenient cover is a file cabinet plastic label holder. The label holder is cut to 1 x 1 inch (25 x 25 mm). The covers will then slip easily over the fiberglass paddle. To give a contrasting appearance, a black rubber alligator clip boot is cut and stretched over the clear plastic.

This type of key arrangement lends itself to a consolidated keyer design. Although an Iambic keyer board was used, the same mechanical arrangements should work with any type of keyer circuit. I hope this description will benefit those contemplating their own keyer design but haven't found a solution to the key packaging problem.

reference

1. Myron Hexter, "The Siamese Paddle," *ARRL Hints and Kinks*, Volume VI, Second Printing, 1959.

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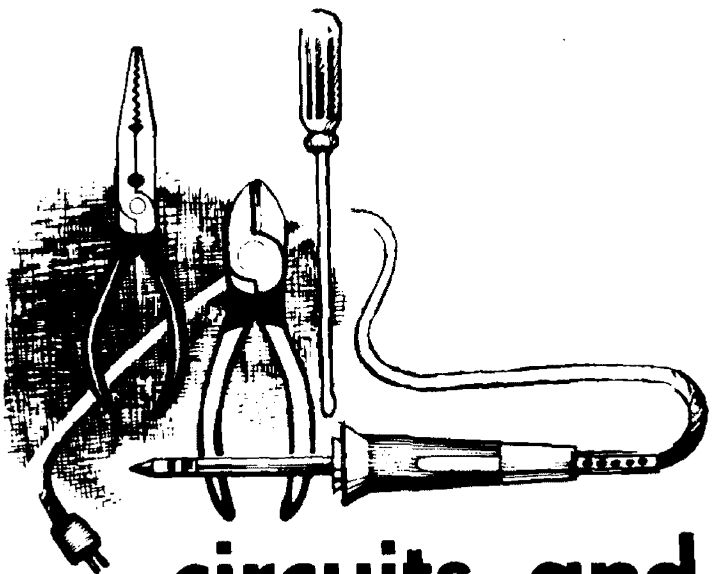


circuits and techniques

ed noll, W3FQJ

winding up the wind

The sun blesses us with warmth and light, each a non-polluting form of abundant energy. By indirect means the sun begets the wind, a conveyer of tremendous energy, since it is the ever-shifting differential heating of the earth's surface by the sun that launches the wind. It blows over the wind-swept plain, races above the mountain tops and dances among the sea cliffs. It also howls and whistles across your back-yard. The conversion of its energy to electricity can easily power your amateur radio station.

How much non-polluting energy is available from the wind? Estimates vary from 20 billion kilowatts capacity at choice sites to a maximum of at least 80 trillion kilowatts to be derived from the winds of the Northern Hemisphere. The peak summer power demand in 1973 was only 344 million kilowatts. In 1970 the peak world demand was 1 billion kilowatts; approximately one-third of this peak capacity was made available in the United States. The world has to struggle to generate this quantity of electrical energy while the vast non-polluting energy sources continue to be ignored – light and heat from the sun, the wind, the tides and geothermal warmth.

Today it is possible to use solar energy to heat your home (with very little augmentation, even in the northern states). You could supply a goodly portion or all of your electrical needs by wind and light energy converters. Would you like to start out on your ham station?

I am doing just this at W3FQJ, struggling toward a hybrid combination that will permit me to use the wind to full-charge batteries while solar cells provide a trickle charge. I am aiming for a 1000-watt capability.

general plan

What are the major units of a wind-to-electrical-energy conversion system? The initial step is the conversion of wind energy to mechanical motion, **fig. 1**. This can be accomplished by a variety of wind blades and rotors such as multi-bladed metal or wooden propellers, sail blades made of light frames covered with material, Princeton and Chalk rotators, Savonius S-rotors, vertical axis rotators, etc. Each has its own particular characteristics and meets specific needs. Various blade types will be described in succeeding columns.

Electrical generators of various types can be used to convert mechanical rotation to electricity. Often an inter-

vening belt and/or gear system makes an appropriate conversion between the rotational speed of the wind-driven rotor and a favorable speed of rotation for the generator. Others are directly driven by wind-driven rotors. Either dc or ac generators can be employed with various voltage and power capabilities as required by the system application. Mechanical arrangements can also be set up to rotate more than one generator. Popular generator-voltage values are 12, 24, 36 and 110 volts. For amateur stations 12-volt systems are likely to be popular because of the availability of low-cost automobile generators and alternators.

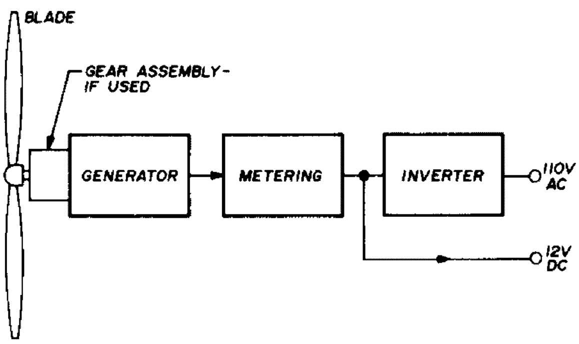


fig. 1. Basic wind generating system. Typical commercial wind generators are shown in the photographs of figs. 2 and 3.

storing electrical energy

There are periods of calm and low wind velocity when generators become inactive. Electrical energy must be stored by batteries for these time intervals. Presently the secondary battery serves as the reservoir of electrical energy. The system must be planned according to the average wind-velocity conditions of an area with an added safety factor that will accommodate a reasonable length wind-quiet period.

Lead-acid batteries are currently the most economical types for medium- or high-powered systems. Other types of secondary batteries are practical for low-power installations. Smaller and better battery systems are inevitable. Despite the debilitating public relation outputs from the current energy empires, *continuous* service can be supplied by solar and wind electrical generating systems.

At the present time high-powered, continuous-service systems are expensive simply because they are not widespread. One can anticipate, therefore, that in the interest of economy, backup systems can incorporate means for using utility services or a fuel-driven generator. Furthermore, in many installations wind generators will be designed to provide only a percentage of the total need. A radio amateur may have as his first objective the self-sufficient powering of his amateur equipment. Later he may wish to incorporate additional electrical needs of his dwelling into the overall system.

electrical energy distribution

Ingenuity in meeting desired requirements is the key word. Presently, for most applications, the most expedient manner of operation is to use the wind generator to charge batteries. In suitable wind and with adequate capacity the output of a wind generator can be made to supply the load as well as to charge batteries. The most popular manner of distribution now is with dc voltages. You will find that for almost any ac motor application you will be able to find a dc replacement. Low-voltage bulbs are readily available for dc light distribution. For those devices and equipment that require 110-volt ac, you can use an inverter. Solid-state inverters with ratings from 50 to 75 watts to as high as five kilowatts are available.

Much modern amateur equipment is designed to operate on 12 volts dc. Fortunately, 12-volt wind generating systems for low and moderate power applications are currently the least expensive. They can also be homebrewed readily at low cost, particularly if you are willing to scrounge in auto junk yards and surplus stores for com-

ponents. Low-powered inverters of low cost can be purchased right out of the electronic parts catalogs. New and rebuilt auto generators and alternators are available from auto stores and catalogs.

fig. 2. A 200-watt wind generator suitable for many amateur radio stations (photo courtesy Wincharger).

reading the wind

The conversion of wind energy to mechanical rotation has every indication of being a complex science. Windmills have been available for generations but have only been conceived and used in rather simple forms. Types of blades and rotors respond differentially to wind. One might be efficient at low velocity; another much more efficient at high velocity. There are now many arrangements under investigation.

In planning a wind generator installation find out as much as you can about your local wind conditions. Such information can be obtained from your local weather bureau or directly from the National Climatic Center, Federal Building, Ashville, North Carolina 28801. A particular interest should be the yearly mean wind speed (average wind velocity).

For most presently available wind generators the minimum average is about 8 mph (12.9 kmh) or more. Average rating of equipment is often based on a velocity of 10 mph (16.1 kmh). In practice, however, the real energy producing winds range between 15-25 mph (24-40 kmh). If these levels are maintained for an average two-day interval out of seven throughout the year you are in a practical situation for a wind-generating system when using a battery-storage plan. If you enjoy a site with wind values better than these, you will have a bonus in extra power.

A small, commercially available and practical 12-volt wind generator, fig. 2, provides approximately 20 kilowatt hours per month in an area of 10 mph (16 kmh) average wind velocity.* If your site averages 16 mph (26 kmh) instead of 10, the same generator has a capability of 35 kWh per month.

Usually the wattage rating of such a generator is based on the minumum wind velocity that will produce maximum output. For example, the unit of fig. 2 produces about 200 to 250 watts output with a wind velocity of 23 mph (37 kmh). Power output levels fall off above this wind speed. In fact, automatic braking systems are included to hold down rotational speed and prevent breakdown from excessive wear.

power calculations

What are your amateur station

*For more information on the Wincharger line of wind-driven electrical generators, write to Mr. Ed Hult, Winco-Dyna Technology Inc., Box 3263, Sioux City, Iowa 51102. For more information on Solar Wind equipment, write to Mr. Henry Clews, Solar Wind, Box 7, Bar Harbor Road, East Holden, Maine 04429.

needs? Go to your log and collect some figures. Here is an example: Assume on transmit that your power demand is 120 watts. This represents a current demand (with a 12-volt supply) of

$$I = \frac{P}{E} = \frac{120}{12} = 10 \text{ amperes}$$

Assume you operate four hours per day and have a transmitter on-time of about 40%. Remember that some current is also drawn on receive and standby but for solid-state equipment this is usually relatively low. Assume a 50% *on time* to compensate for this additional current. (This approximates a two-hour period of operation drawing 120 watts continuously.) Daily consumption then approximates 20 ampere-hours (10 x 0.5 x 4). Monthly consumption would be 600 ampere-hours (30 x 20).

A suitable heavy-duty 120- to 150-ampere-hour 12-volt battery would adequately supply these needs. A second battery would give plenty of backup power as it could be kept on charge while the first battery is in operating position. Furthermore, a solar panel to provide a trickle charge to the battery in use will give you an additional safety factor.

How would you fare under 24 hours of continuous operation? This would correspond to a time period of 12 hours of continuous 120-watt demand. A very maximum consumption would be 240 ampere-hours. You even have some power in reserve for a very extended period of operation. You would have no trouble operating a 180- to 200-watt PEP solid-state transceiver under the above circumstances. In fact, you would have power to spare.

What would be your monthly consumption of power based on the above average conditions? Your daily use of electricity would be 0.24 kilowatt-hour (120 x 2). This would be 7.2 kWh per month (30 x 0.24). Based on the rating of the wind generator of fig. 2 you again would have extra power because its rating at an average wind speed of 10 mph (16 kmh) is approximately 20 kWh. Therefore, you have extra power to operate additional equipment in the ham shack, operate more hours or use a bit more power.

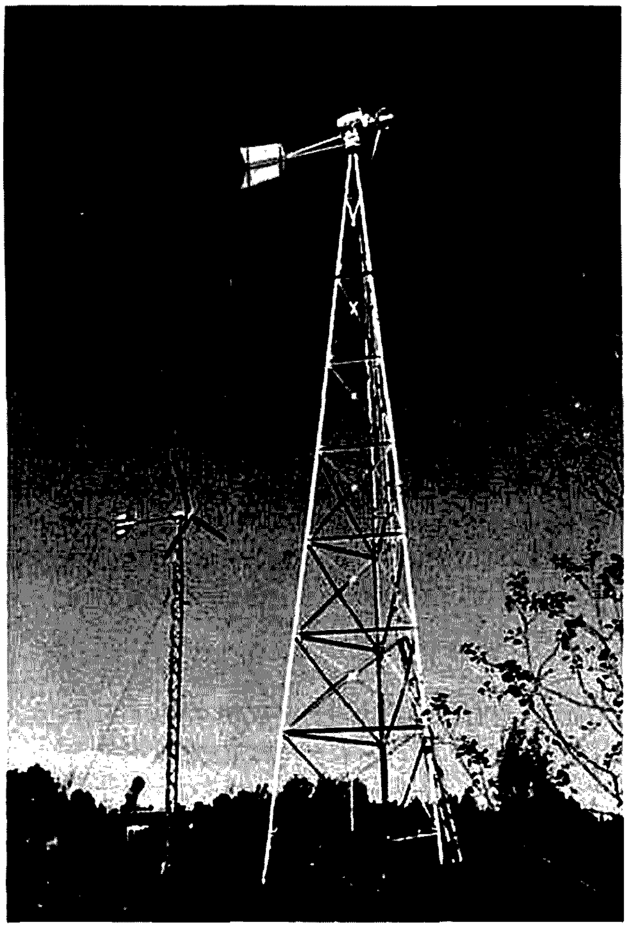
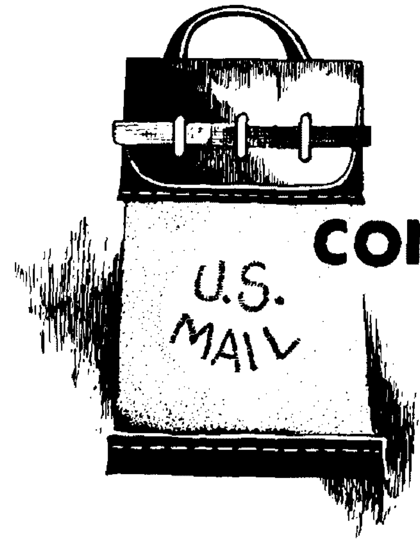


fig. 3. This 6-kW wind generator (foreground) and 2-kW unit (to the rear) supply power to a business (photo courtesy Solar Wind).

These calculations have been very conservative and indicate how the smallest of wind-generating systems can make a modest ham station self-sufficient electrically. Knowing the ingenuity of radio hams, each, according to his operating habits, will find the most efficient use of the generated power. In fact, some of the power will probably be diverted to other services in the house. As the lament says, ". . . the answer is blowing in the wind."

ham radio



comments

5/8-wavelength antennas

Dear HR:

Having built and used both quarter- and 5/8-wavelength ground-plane and mobile antennas on two meters, I was surprised at the article in the May, 1974, issue of *ham radio*. I have had very good results using the 5/8-wavelength antenna, and KØDOK's article spurred me to do some of my own testing to see the difference for myself.

I decided that my Heathkit HW202 with a step attenuator between the antenna and transceiver would make a good rf micro-voltmeter with which to make the measurements. The antenna is a portable ground plane consisting of three 5-foot sections of telescoping TV masting, guyed and set on a pin so the antenna could be lowered to change the driven elements and be put back in exactly the same position (I used nylon guy ropes which had enough stretch so it was unnecessary to change the guy ropes between antenna changes). The ground radials are mounted to a steel angle bracket with wing nuts for portability. The antenna is attached to a bracket with a PL-259 plug which connects to a SO-239 jack on the mounting plate.

The 5/8-wavelength element consists of a series matching inductor to the 5/8-wavelength radiator with a vswr of 1.7:1. The quarter-wavelength radiator has a vswr of 1.2:1. Both antennas were mounted on the same 15-foot mast.

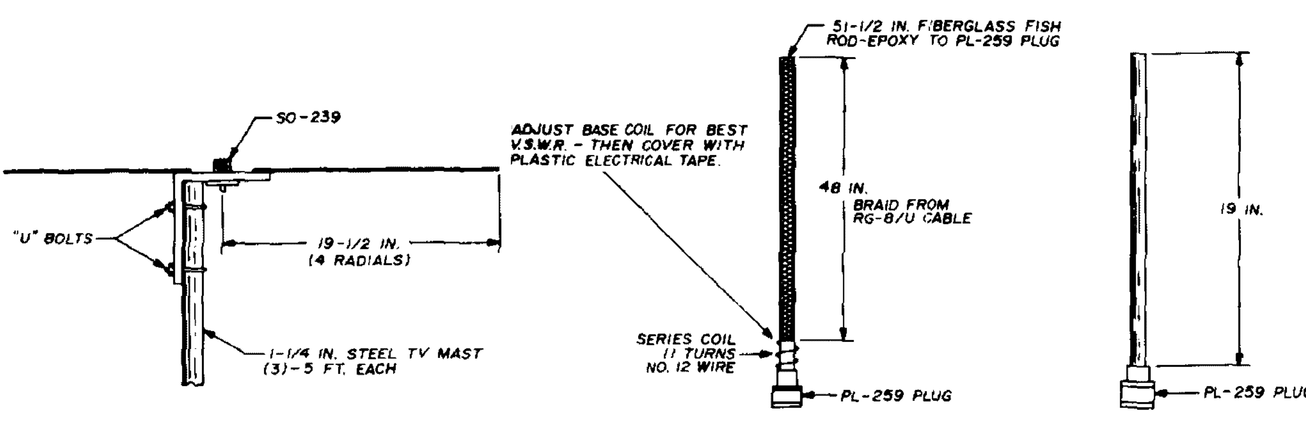


fig. 1. Details of the antennas, ground radials and mounting bracket used in gain comparisons.

The measurement plan was to use received signals, attenuate the signal to a mark on the S-meter in the transceiver, note the attenuation, take the antenna down, change the driven element, put the antenna back in place and again attenuate the signal to the same meter mark, record the reading and note the results. The step attenuator used has a range of zero to 101 dB in 1-dB steps. The results are shown in table 1.

Stations of different distances were used to take into account the different angle of radiations between the two antennas. Every effort was made to replace the antenna in exactly the same spot as before so that the tests were all based on a very minimum of variables.

In conclusion, the data in table 1 shows that the 5/8-wavelength antenna has 3-dB gain over the ¼-wavelength antenna. The antennas used for making these tests are shown in fig. 1.

John H. Pearson, K2GVP
Baldwinsville, New York

I can find no obvious fault with Mr. Pearson's measurement technique although he has used a crude method at best. However, the averaging of a number of results of this type of measurement should yield a meaningful result. That is in effect what Mr. Pearson has done although his measurement resolution of one dB and the small number of samples (three) makes it hard to find the true average. I would conclude from the data however, that Mr. Pearson did indeed achieve gain from the 5/8-wavelength radiator.

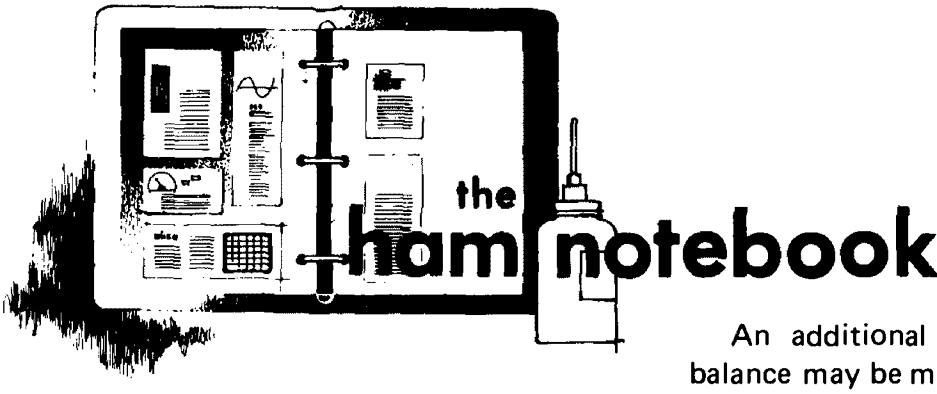
As I pointed out in my article, obtaining this performance from a monopole depends upon creating an image element, hopefully from a reflection in the ground plane. I suspect that what Mr. Pearson has actually managed to do is to excite sufficient currents along his 15-foot support mast to, in effect, create an image radiating element. There is at least one commercial antenna (the Ringo*) which appears to depend on this as it does not even include ground-plane rods. From a designer's point of view this is a dangerous approach because the designer has no control over the antenna mounting structure and very few amateur installations will probably wind up having the antenna atop a long straight conducting pole. More likely the antenna will be set just above a 20-meter beam or even side-bracketed to a tower.*

It would, of course, be very interesting if we could see a vertical plane radiation pattern of each of Mr. Pearson's antennas to see if a correspondence exists to fig. 8 *and* 9 *of my article. Even given a well equipped antenna test range, that measurement would be a tricky one to make. On several occasions I have attempted range measurements at 150 MHz and have always been plagued by ground reflections, especially when tipping the antenna on its side so that it could be rotated for a pattern in the plane of the monopole. That is why the model work reported in my article was conducted at 1000 MHz where reflections are far easier to control.*

Paul E. Meyer, KØDOK

table 1. Gain comparison of 5/8- and 1/4-wavelength two-meter groundplane antennas with three different stations. Measurement technique is discussed in letter.

	distance	1/4-wave antenna	5/8-wave antenna	gain
Station 1, Base Station	11 miles	18 dB	21 dB	+3 dB
Station 2, Repeater	21 miles	31 dB	34 dB	+3 dB
Station 3, Repeater	41 miles	4 dB	7 dB	+3 dB



better balancing of the Heath HM-2102 wattmeter

After wiring the Heath vhf wattmeter very carefully it may be rather disappointing to note that anything below a swr of 1.5:1 is apparently considered normal in balancing the bridge, and as designed the bridge very often cannot be nulled much below a 1.5:1 reading. Of course, it is assumed a good 50 ohm load is used at the nulling frequency, such as a Bird dummy load. However, if after following the nulling procedure per the Heath manual no good null can be obtained, the following change may be considered.

Exchange C3 (7.7 pF) and C16 (10 pF) so C16 becomes a capacitor of 7.7 pF. In parallel with C16 mount one of the 1/4-inch (6-mm) diameter miniature ceramic trimmer capacitors with a range of about 1.5 – 7 pF. If this is not available mount only a miniature trimmer with a range of about 5 – 15 pF. The intention here is to make C16 variable around 10 pF. By alternately adjusting C4 and C16 in the nulling procedure an excellent null can be obtained, much better than is possible with a fixed value of C16. Or C3 should have a greater value, in many cases it seems to, and this can be judged from the position of the rotor of C4 when this capacitor is adjusted for maximum null before modification. If C4 is fully closed C3 should be increased in value.

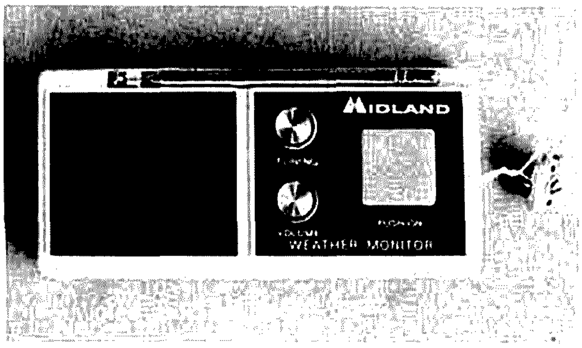
An additional check for sensitivity balance may be made as follows. Set the sensitivity control to maximum (minimum resistance) and with no load on the antenna side of the bridge feed just enough rf power into the bridge for a full-scale deflection on the meter in the forward position. Switching now to the reference position should give the same indication on the meter. *If there is* a difference of more than 10%, a correction can be made by adding additional resistance in series with the higher reading line at either point B or C on the PC board – start with about 470 ohms. This test can only be done with vhf rigs able to operate without a load (such as the Heath HW-202). If the rf output level of the rig cannot be reduced as needed, the bridge's sensitivity control may be used instead but this will reduce the accuracy of the test, of course.

Bob Fransen, VE6RF

retune weather monitor receiver to two-meter fm

A quick glance through any of several mail-order catalogs will reveal that there are inexpensive transistor receivers available for use as "weather monitors." These fixed-frequency receivers are designed to receive vhf-fm continuous weather information transmissions from National Oceanic and Atmospheric Administration (NOAA) stations on either 162.40 or 162.55 MHz. The broadcasts, a part of the National Weather Service, are intended to provide

local weather information in the public interest, especially where natural hazards are involved. This is a great idea, and many people have bought these units with thoughts of keeping track of hurricanes, receiving tornado warnings, and for assistance in planning weekend boat trips.



This particular model receiver already had a very small trimmer "Tuning" control to provide reception of either 162.40 MHz or 162.55 MHz.

However, quite a few would-be listeners have found that these units lack the sensitivity to detect the low power government transmitters at their homes and consequently the fixed-tuned receivers are useless to them. This is understandable when you compare the fine print of the receiver specs in the catalog and note a 7.5-μV sensitivity rating – and then remember the NOAA recommendation of a receiver with capabilities of 1.2 μV for 40 miles from the station; 0.9 μV for 50 miles and 0.6 μV for 60 miles.

Now, if you know of someone who has one of these little boxes and is about to throw it out, you might casually mumble that you, "could use the radio for parts" and, with luck, carry it back to your shack. Upon inspection, you will find that you possess a 9-transistor (typically) fm receiver which, with the addition of one capacitor, can be tuned to any frequency you want, such as the local repeater. Mine is set for 146.94 MHz.

Simply remove the unit's cover, locate the local oscillator circuit, and place a padder capacitor across its tuned network. As shown in the photograph, I used a small variable trimmer and placed it outside the receiver. With this arrangement I was able to tune down into the commercial fm band (88 to 108 MHz), aircraft communications frequencies (108 to 136 MHz) and, of course, the entire amateur 2-meter band. When a variable capacitor has been found that allows coverage of the desired range of frequencies, it may be permanently installed within the receiver.

Kent Mitchell, W3WTO

cmos keying circuits

Shown in fig. 1 are two alternate keying circuits which may be used with the cmos electronic keyer which appeared in the June, 1974, issue of *ham radio*. The circuit in fig. 1A is designed for grid-block keying and will key up to –140 volts, maximum. The circuit in fig. **1B** is designed for keying solid-state QRP transmitters.

Jim Pollock, WB2DFA

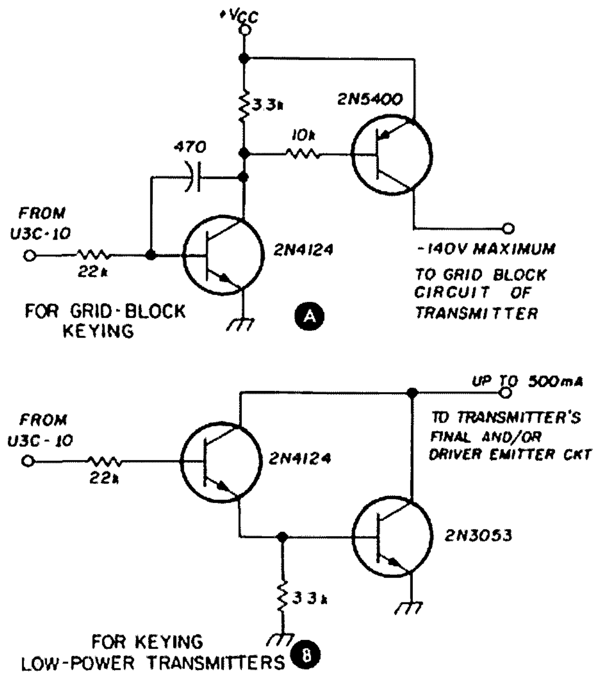
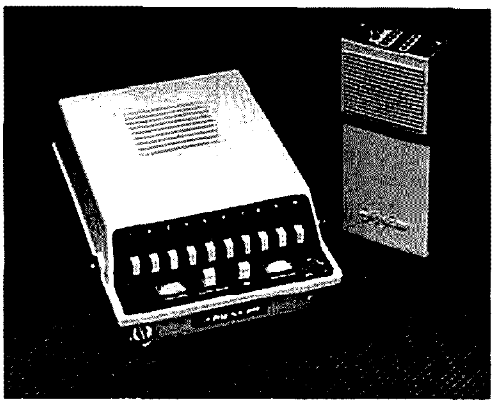


fig. 1. Alternate keying circuits for use with the cmos electronic keyer.

new products

monitoring scanner receivers



The Hy-Gain Electronics Corporation has introduced two new vhf scanner receivers, the Hy-Scan 10 for ten fm channels, and the pocket model Hy-Scan 4 for monitoring four channels. The Hy-Scan 4 is the second generation of Hy-Gain's popular Pocket Scanner, and includes several new features including individual channel lock-out, continuous volume and squelch controls, external antenna jack, earphone jack and an auto-manual scan switch. All of this is built into the small size of 5.5x2.5x1.2 inches (14.0x6.4x3.0 cm). The Hy-Scan 4 operates on four AA batteries and uses standard 10.7-MHz i-f scanner crystals. Models are available for 30-50 MHz, 150-170 MHz and 450-470 MHz.

The Hy-Scan 10 is the big brother to the Hy-Scan 4 Pocket Scanner and features a rugged molded cycolac case with carrying handle with a flip-top which allows instant access for changing the plug-in rf modules in the single-band models (a three-band model is also available). The rf modules allow the choice of 30-50 MHz, 150-170 MHz or 450-470 MHz operation. Function controls include LED channel indicators, individual lockouts for each channel and automatic or manual scan select switch. The automatic battery charger and hide-away telescopic antenna are built in. Measuring 2.5x5.5x8 inches (6.4x14.0x20.3 cm), the Hy-Scan 10 has jacks for earphone and an external antenna, and uses standard 10.7-MHz i-f crystals.

For more information on either the ten-channel Hy-Scan 10 or the Hy-Scan 4 Pocket Scanner, write to Hy-Gain Electronics Corporation, 8601 Northeast Highway 6, Lincoln, Nebraska 68505, or use *check-off* on page 94.

test equipment catalog

A six-page condensed catalog featuring Eico's broad line of electronic test and measuring instruments for laboratories, industry and amateurs is now available from Eico Electronic Instrument Company. The all new catalog features the most popular units in Eico's line of over 100 electronic kits and factory-assembled instruments, including oscilloscopes, VTVMs, VOMs, signal generators, tube/transistor testers,

More Details? CHECK—OFF Page 94

power supplies, as well as the Truvohm line of multimeters. For your copy of this new catalog, write to Eico Electronic Instrument Co., Inc., 283 Malta Street, Brooklyn, New York 11207, or use *check-off* on page 94.

TTL cookbook

This new book by Donald Lancaster covers practically every aspect of TTL logic, shows you what TTL is, how it works, and how to use it in your own circuits and projects. In the first chapter the basics of TTL are given: what it is, how to interconnect it, how to power it, and so on. Chapter 2 is a catalog of TTL devices, giving physical and electrical specifications of all the devices mentioned in the book. Logic is covered in Chapter 3, starting with the usual basics, then going to more advanced logic designs. Particular attention is given to showing how TTL yields single-package solutions to traditionally difficult problems.

Gate and timer circuits are discussed in Chapter 4. Some practical applications are given including controlled oscillators, two-tone alarms, digital capacitance measurement, frequency meters, digital thermometers and others.

Succeeding chapters take up clocked logic, JK and D-type flip-flops and applications, counters and counting techniques, shift-register circuits, noise generators and rate multipliers. The final chapter discusses a number of practical applications, including digital counter and display systems, events counter, electronic stopwatch, digital voltmeter, digital tachometer, and other digital instruments. The author suggests several TTL projects that the reader may wish to try, relying on his own resources.

This book, by the same author of the very popular RTL *Cookbook*, is highly recommended to anyone who wants to

know more about TTL ICs and how to use them. Published by Howard Sams & Company, 336 pages, soft-bound, $8.95 from Ham Radio Books, Greenville, New Hampshire 03048.

swan ssb transceiver

The new Swan 700CX Transceiver, priced under $1 per watt, has a 700-watt PEP input rating on single-sideband, power enough to punch through QRM without need for an accessory linear amplifier. Frequency ranges span the 10-, 15-, 20-, 40- and 80-meter bands with extended frequency coverage of up to ten channels for MARS operation (available with use of an optional plug-in crystal-controlled oscillator).

Operation includes selectable USB, LSB, a-m and CW modes with standard i-f filtering provided through the use of a 5.5-MHz crystal filter with a 2.7-kHz bandwidth having a 1.7 shape factor and ultimate rejection in excess of 100 dB. Included in the new 700CX is a built-in, selectable 25- or 100-kHz crystal calibrator.

Other optional accessories include power supplies for dc or ac operation, a 2000-watt linear amplifier, external vfo, a super-selective i-f filter kit, plug-in vox, phone patch and mobile mounting kits. Price of the basic transceiver is $599.95. For more information, write to Swan Electronics Corporation, a subsidiary of Cubic Corporation, 305 Airport Road, Oceanside, California 92054, or use *check-off* on page 94.

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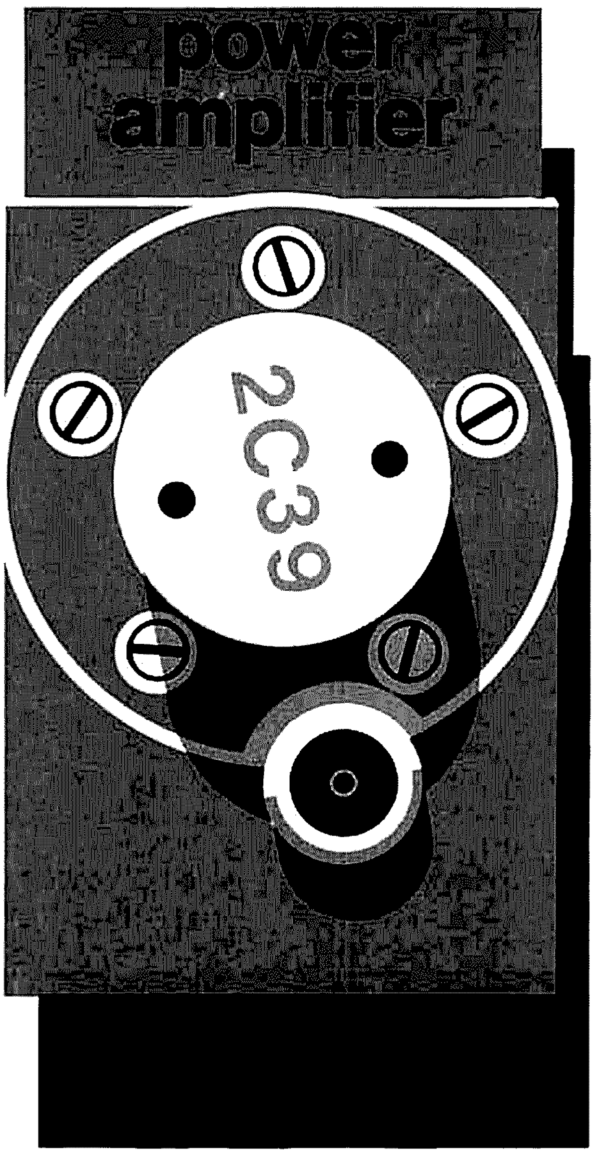
FEBRUARY 1975

2304-MHz



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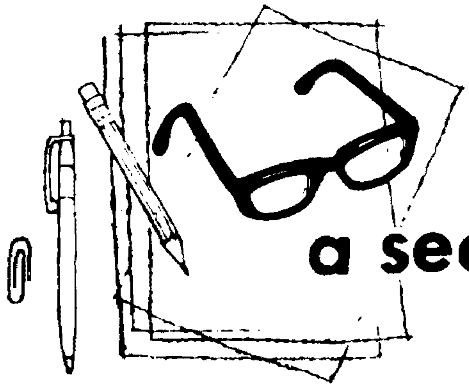


a second look

by jim fisk

The FCC finally released its long-rumored "amateur restructuring" proposal in mid December which, among other things, would create two new amateur license classes and re-arrange the frequency, power and emission privileges. And, as so often happens, early truncated reports of Docket 20282 were rushed into print so quickly that they glossed over some important details — details which, when presented in the proper light, would clear up most of the misunderstandings which many amateurs apparently have about the sweeping new proposals outlined in the 29-page Docket.

First, and most important, if the proposal is adopted *presently* licensed amateurs would gain much more than they lose. True, separate high-frequency and vhf licenses are proposed, but *present* General and Advanced Class licensees may obtain their counterpart vhf licenses simply by request. In addition, Advanced Class licensees would gain use of the high-frequency radiotelephone segments now reserved for the Amateur Extra Class as well as a maximum power limitation of 2000 watts PEP *output*, a substantial increase. Advanced Class licensees would lose their operating privileges above 29 MHz (including 29.0 to 29.7 MHz) but only until they applied for and received their Experimenter license, the new vhf counterpart to the Advanced Class which carries all operating privileges above 29 MHz.

General Class licensees would retain their present operating privileges below 29 MHz with a permissable maximum power limitation of 500 watts PEP *output*. Considering the average efficiency of rf power amplifiers, this represents only a modest decrease from the present power level. To regain their vhf privileges above 50 MHz Generals would have to apply for a separate Technician Class license — no additional examinations would be required. *New* licensees would, however, be required to pass separate examinations for each class of license they desired.

Under the new proposal the Novice Class license would be renewable for five-year terms as the other classes are now, and the maximum power limitation would be 250 watts *input*, another substantial increase. Since Novice licensees are limited to CW operation, the Commission felt that the traditional "voltage times current" measurement of input power was still appropriate; such is not the case with other, more advanced modes such as single sideband and slow-scan television. Under the new proposal Technician Class licensees (or other vhf licensees, for that matter) could also hold the Novice Class, an option not now available.

The newest, and in some ways most exciting, proposal contained in Docket 20282 is the Communicator Class license — a code-free amateur license which would offer use of all amateur frequencies above 144 MHz, F3 emission only. The size of the Amateur Radio Service has declined measurably in recent years and the new Communicator Class should do much to start our ranks growing again. Some amateurs are opposed to the idea of a code-free license simply as a matter of tradition,

(continued on page 43)

presstop

OSCAR 7 ORBITAL PERIOD has been determined to be 114.945 minutes, sufficiently shorter than initially reported to account for the incorrect arrival times in early orbital predictions. The error was NORAD's -- their tracking station had confused the rocket's second stage with OSCAR 7! The corrected figure for earth movement between orbits is now 28.74° at the equator.

Complete Updated Orbital Data for OSCAR 6 and OSCAR 7 will be provided monthly as an added slip-in sheet to HR Report. Copies of these predictions are available to all interested readers upon receipt of an SASE (one SASE for each month).

OSCAR 6 Is Being Abused, and AMSAT officials are concerned. It is important to 6's future that it be used only during the scheduled on periods, even though it may be found on at other times. Limit your use of OSCAR 6 to Monday, Thursday and Friday (GMT) for afternoon or evening contacts, plus Sunday mornings.

MORE MOONBOUNCE TESTS planned for February from WA6LET, using Stanford's big dish on 144.190 and 432.190 and listening down about 90 kHz. Operating schedule will be 0500-1000Z February 2 and 0000-0500Z February 23, with WA6LET transmitting the first half of each minute and listening for callers the second half. For further details write Victor R. Frank, Stanford Research Institute, Bldg. 320A, 333 Ravenswood Ave., Menlo Park, California 94025.

FCC POLICY REGARDING CODED TRANSMISSIONS (Presstop, January) has been receiving further study at the Commission. It appears now that control signal transmissions will not be involved, and only telemetered data transmissions would require disclosure. Until specific details of a procedure for advising the FCC of your telemetry transmissions have been worked out, the requirement for advising local FCC offices has been withdrawn. Full details on the procedure will be provided when they become available.

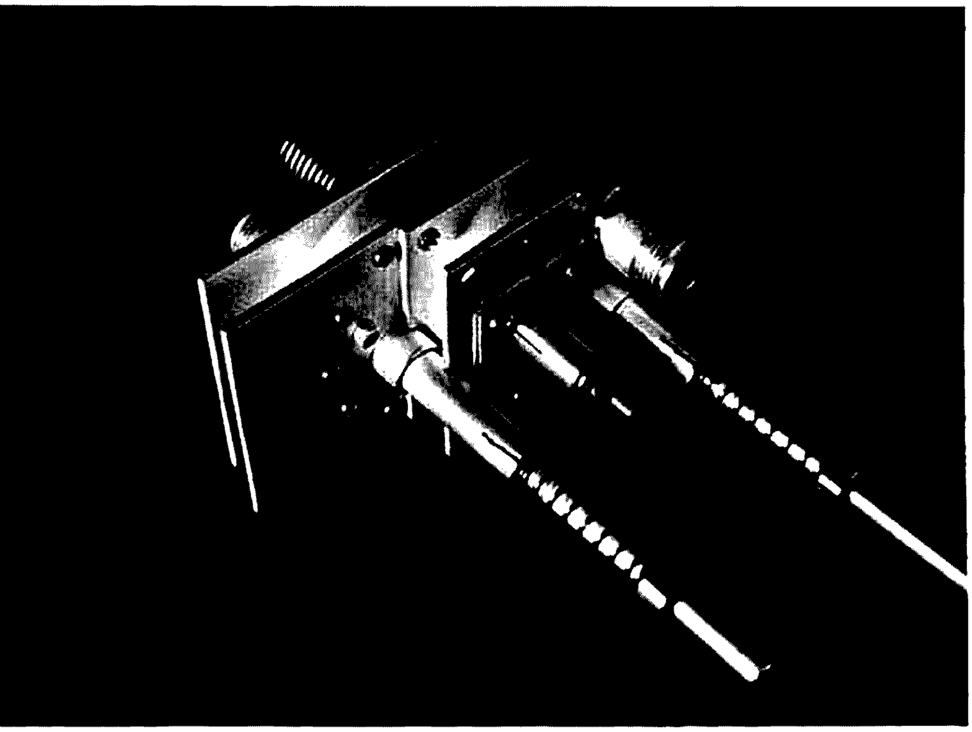
REPEATERS STILL OPERATING WITHOUT WR CALLS are about to get the axe -- FCC believes it has now processed all on-hand applications for "grandfathered" repeaters, so if you are still waiting you'd better check with the FCC in Washington. FCC records show that more than forty early repeater applications that were returned to the applicants for further information have never come back to the Commission for action.

WWV HAS ADDED USEFUL PROPAGATION INFO to its 14-minute-after-the-hour current radio conditions report. "K Index" refers to current geomagnetic conditions: 1-2 is okay but watch out for 3 or 4. The-higher-the-better is the rule for solar flux. Anyone who plots this information on an hour-by-hour basis and correlates it to conditions on the various bands might find he had developed a nice competitive weapon for stalking DX!

KLM'S AD FOR NEW 2-METER TRANSCEIVER will break this month -- it's a 10-watt PEP frequency-synthesized CW/SSB rig ideal for two-meter DXing and OSCAR work, according to Mike Stahl, K6MYC. Called the ECHO II, this diminutive 8-pound package features a built-in noise blanker, CW break-in, and receiver RIT and comes set up for 145.0-145.23 and 145.77-146.0 MHz. Price is $389, and KLM plans some nice package deals with solid-state linear amplifiers and antennas. For more info write to KLM Electronics, Dept. H, 1600 Decker Avenue, San Martin, California 95046.

CENTENNIAL CALL SIGN ideas for 1976 still wanted by FCC. Ground rules are that special prefixes must be "self-assignable" and non-ambiguous. In addition to present W/WA-WZ and K/KA-KZ blocks, AA-AL, N and NA-NV may also be used. The catches are to avoid any presently used prefixes (KA, K8, WP and WL, for example) and those likely to show up as part of the restructuring effort. Send your ideas to Prose Walker at the FCC now as work is expected to begin on the project almost immediately.

More Details? CHECK—OFF Page 94



2304-MHz power amplifier

Complete construction details for a single-tube 2304-MHz amplifier that delivers 30 watts output and up to 13-dB gain

Norman J. Foot, WA9HUV, Elmhurst, Illinois 60216

Of the many varieties of special-purpose tubes which give good performance on the uhf bands and above, the 2C39 remains as the outstanding candidate for amateur use because it is readily available and the price is right. The recent onslaught toward achieving high power on 1296 MHz is directly attributable to the use of the 2C39 and its derivatives. Numerous single-tube designs were initially reported in the amateur publications, followed by multiple tube designs[1,2] and, finally, the popular octet of 2C39s described by Peter Laakmann[3] in 1968.

More recently, interest in 2304 MHz has been growing, and a few pioneering efforts to achieve reasonable rf power without the help of high power klystrons have been reported.

This article is concerned with the design of a 2304-MHz power amplifier which uses a single 2C39 tube. Calorim-

eter measurements show that 30 watts of rf power output at 25% plate efficiency are achievable with a nominal power gain of 13 dB. The design is rugged and performance is stable. All of the parts for this power amplifier can be made with hand tools with the exception of the cavity rings; these should be cut and faced off in a lathe. The filament and cathode socket parts were obtained from surplus 2C39 amplifiers.

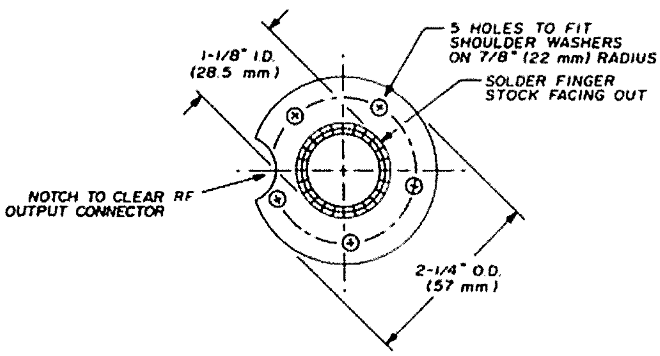


fig. 1. Plate ring for the 2304-MHz power amplifier. This ring is mounted on the output plate (fig. 2) with a 10 mil (0.25 mm) Teflon insulating sheet. Material is 1/8" (3mm) brass.

The only parts requiring soldering with a torch are the finger stock, the tuning bushings and the type-N input and output connectors. All other parts are screwed together, including the cavity rings. Like its 1296-MHz counterpart, the 2304-MHz 2C39 amplifier uses cavity resonators in both the cathode and plate circuits. Both cavities are 3/8-inch (9.5-mm) long. The cathode cavity has a 2-inch (51mm) inside diameter while the inside diameter of the plate cavity is 1-3/4 inch (44.5mm).

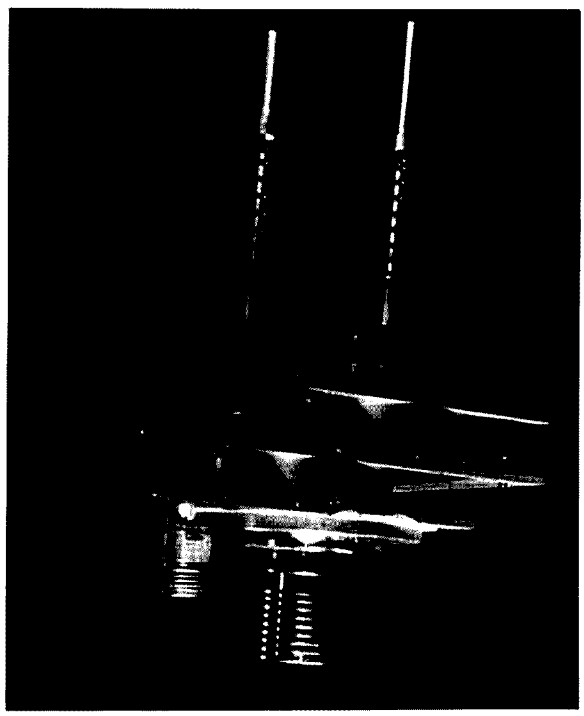
Although the cavity volumes are relatively small, there is room enough for the 2C39, a piston tuning capacitor and a type-N coaxial connector if the parts are positioned as illustrated. By necessity the 2C39 is located very close to the edge of the cavity. This physical constraint is fortuitous however, since the input and output impedances of the 2C39 are lower at 2304 MHz than at 1296 MHz. Adequate cavity coupling is achieved when the tube is mounted close to the cavity wall as shown in the photographs.

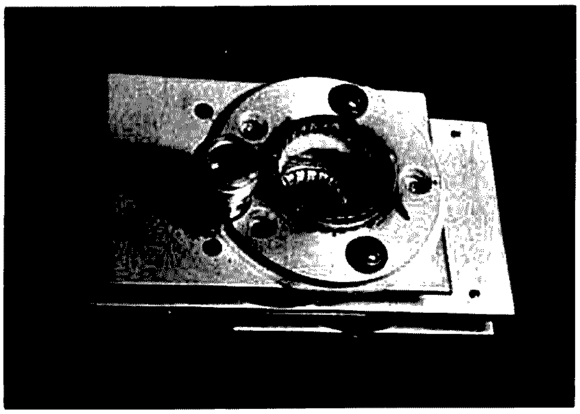
The grid cavity is made of 2¼-inch (57mm) OD brass tube with a 2-inch (51mm) ID. The 1/8-inch (3mm) wall thickness can be drilled and tapped for 2-56 screws if reasonable care is taken. The original 2304-MHz 2C39 amplifier I built also used 2-inch (51mm) ID tubing for the plate cavity but a 2-inch (51mm) OD ring, 1-3/4-inch (44.5mm) ID, was slipped inside of the cavity to bring the resonant frequency up to 2304 MHz. If desired, the plate cavity ring can be a single unit, 2-1/4 inch (57mm) OD and 1-3/4 inch (44.5mm) ID.

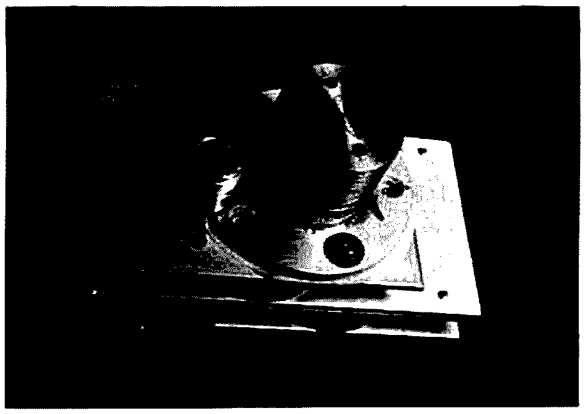
plate assembly

The plate assembly consists of the 1-3/4-inch (44.5mm) ID plate cavity (fig.

Side view of the 30-watt 2304-MHz amplifier showing the 2C39 and output connector (top), plate cavity, cathode cavity and tuning pistons.







View from the plate output side of the amplifier, with and without and 2C39 installed.

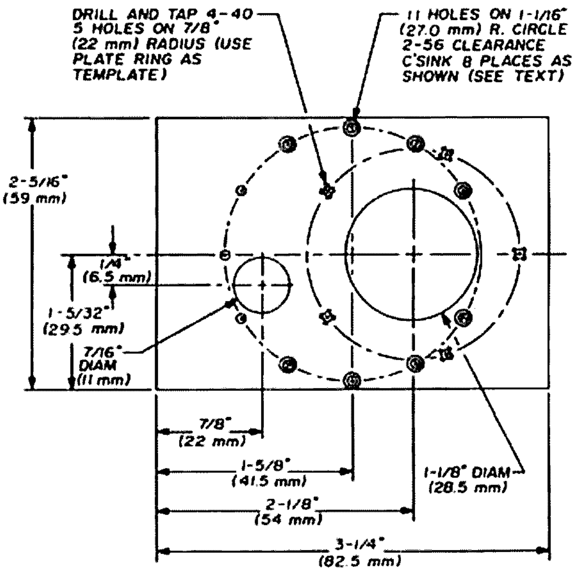


fig. 2. Dimensions for the output plate (outside face). Material is 0.093" (2.5mm) brass.

1) sandwiched between the output plate (fig. 2) and the grid partition (fig. 3). The plate ring with its finger stock is mounted on the outside of the output plate with a 10 mil (0.010 inch or 0.25mm) Teflon insulating sheet. The 11 holes on the output plate are drilled 30° apart on a 1-1/16-inch (27mm) radius circle. Eight of these holes are countersunk for flat-head 2-56 screws so that the plate ring will mount flush. The output plate is used, in turn, as a template for locating the tapped holes in the plate cavity. Fig. 4 shows the plate cavity rings mounted on the output plate. Note that the inner ring is split to clear the rf output coupler.

The type-N output connector, which is made from a UG-58A chassis connector, is soldered in the 7/16-inch (11mm) hole on the otuput plate. The square mounting flange of the UG-58A is first cut off with a hack saw and then the barrel of the fitting is filed smooth. The mounting hole should be drilled undersize and reamed from the outside to provide a slightly tapered hole which will provide a force fit with the connector. When assembled, the Teflon part of the connector should be flush with the

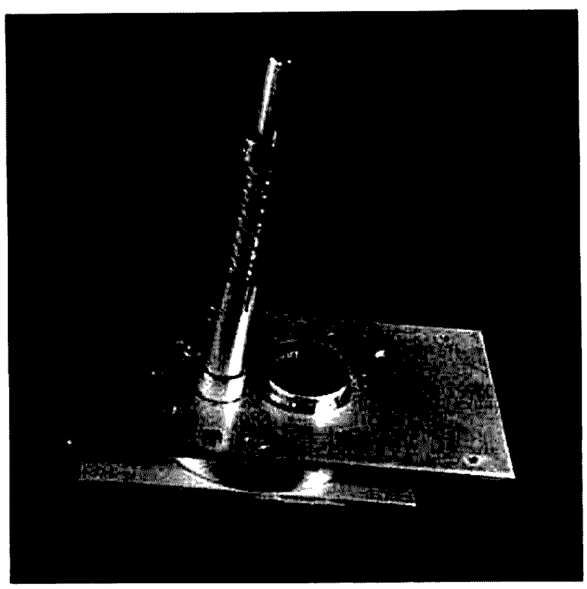


Plate assembly before installation of the cathode cavity.

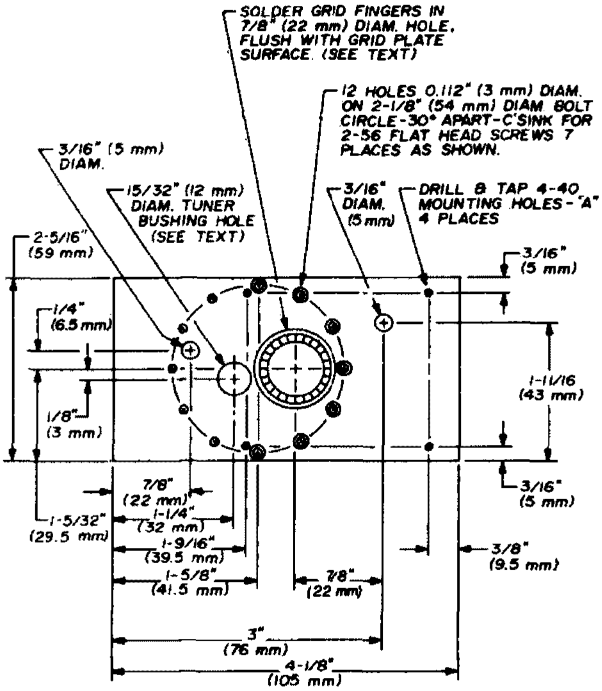


fig. 3. Grid partition for the 2304 amplifier. Material is 0.093" (2.5mm) brass.

inside surface of the output plate. Use a propane torch to provide sufficient heat for soldering. Bring the output plate up to temperature evenly, and avoid prolonged application of the flame directly on the type-N fitting.

The 15/32-inch (12mm) tuner bushing hole in the grid partition (fig. 3) should be drilled undersize, and reamed out to provide a force fit with the bushing. The tuner bushing (fig. 5), extends 1/8 inch (3mm) into the plate cavity when pushed down against its shoulder.

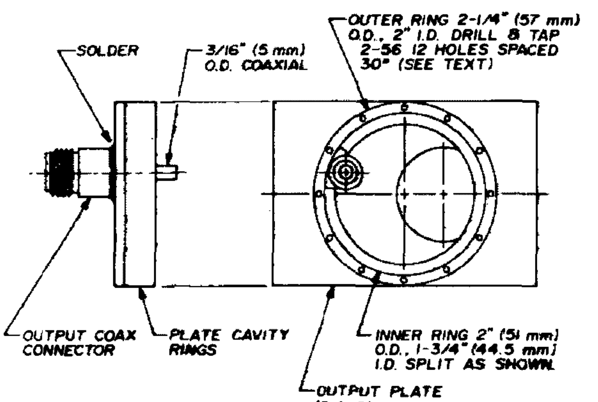


fig. 4. Plate circuit ring assembly (inside view on right).

Before soldering the grid finger stock and the tuner bushing to the grid partition, lay a 9x11-inch (23x28cm) sheet of fine emery cloth face up on a flat metal surface and sand the grid partition flat by moving the metal plate back and forth over the emery cloth. Then polish the surface of the plate with fine steel wool. The other plates should be treated in a similar manner.

The grid finger stock and the tuner bushing are soldered to the grid partition at the same time. The finger stock should be flush with the grid partition on the side facing into the plate cavity. When so located, the grid finger stock exerts a force which pulls the 2C39 into

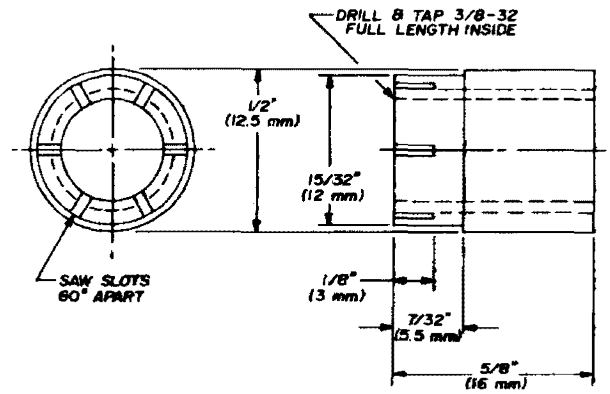


fig. 5. Tuner bushing (two required). Material is ½" (12.5mm) brass tube.

position so that the surface of the grid next to the ceramic insulation of the tube is held against the grid partition. This is critically related to the resonance frequency of the plate cavity. To achieve proper grid finger stock position, lay the grid partition face down on a flat metal surface, insert the finger stock flush against the metal surface, and solder from the rear side using a propane torch.

Before drilling the 12 holes on the grid partition for the plate cavity, mount the plate cavity and the Teflon insulated plate ring to the output plate and insert

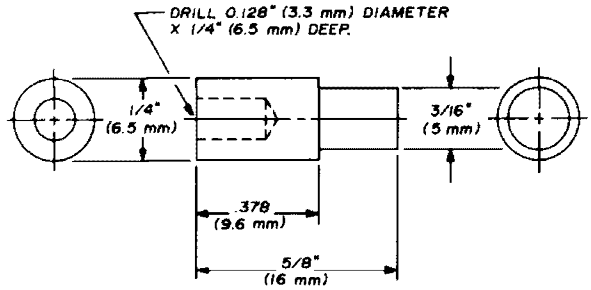


fig. 6. Coaxial couplers. Two are required, one for the input, the other for the output. Material is ¼" (6.5mm) brass rod.

the 2C39. Next, slip the grid partition into place, allowing the grid finger stock to hold the assembly together. Align the edges of the output plate and grid partition so they are parallel. Then, using a scriber, carefully mark the outer edge of the plate cavity on the inside of the grid partition. This will identify the exact position for the plate cavity and will help to properly locate the 12 mounting holes. This is an important step in fabrication since it is essential that the grid finger stock be perfectly aligned with the plate finger stock.

After drilling these 12 holes, reassemble the parts and mark the locations of the 12 holes on the plate cavity. Carefully drill and tap each hole approximately 3/16-inch (5mm) deep. Use 1/4-inch (6.5mm) long 2-56 stainless steel, binder-head screws to attach the grid partition to the cavity.

The 3/16-inch (5mm) hole in the grid partition is a clearance hole for the output coupler (fig. 6). The center conductor of the N connector is fitted with a 1/4-inch (6.5mm) diameter brass rod turned down on one end to 3/16 inch (5mm) to slip-fit into the 3/16-inch hole in the grid partition. The length of the 1/4-inch diameter portion of the coupler should be made a few thousandths of an inch longer than 3/8 inch (approximately 0.378" or 9.6mm) so its shoulder will bear against the grid partition wall when the amplifier is assembled. Before reassembling the output plate circuit, insert the tuning piston shaft (fig. 7) through the bushing from inside the cavity and thread the piston into place.

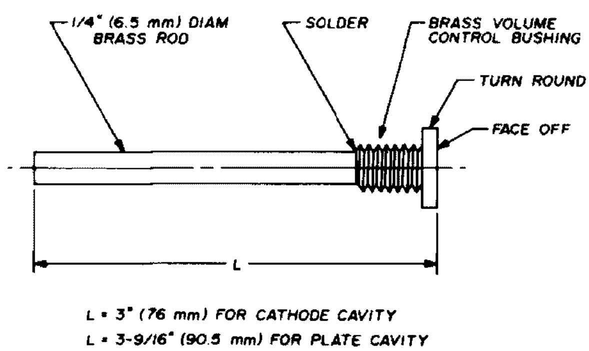
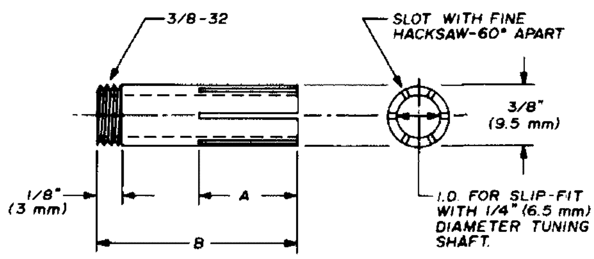


fig. 7. Tuning piston (two required).



	A	B
PLATE SLEEVE	5/8" (16 mm)	1-1/4" (32 mm)
CATHODE SLEEVE	3/8" (9.5 mm)	3/4" (19 mm)

fig. 8. Tuning sleeves. One is required for each of the tuning pistons in the plate and cathode circuits. Material is threaded brass tubing (see text).

tuning pistons

The tuning pistons are made from 3/8-inch (9.5mm) brass bushings from old volume controls and 1/4-inch (6.5mm) brass rods. Insert the rod into the bushing as shown and solder the two together. Clamp an electric drill in a vise between two blocks of wood and mount the 1/4-inch (6.5mm) tuning piston shaft in the chuck. Then, using the hand drill as a lathe, file the hexagonal surface of the bushing round. Also, file the end of the assembly smooth and true.

The tuning pistons are screwed into the tuner bushing from inside of each

cavity. The tuning sleeves (fig. 8) are then slipped over the 1/4-inch (6.5mm) tuning shafts from the outside, and screwed into the tuner bushings. These sleeves provide the necessary mechanical stability and also serve as rf chokes for the tuning pistons; the amplifier should not be operated without them.

If brass tubing with 3/8-32 threads on the inside is not available, tap 13/32-inch (10.5mm) ID brass tubing to a depth of 5/16 inch (8mm) from each end as shown in fig. 5. Turn down the shoulder on one end by using your hand drill as a lathe. One of the tuning pistons can be used as a jig to hold the tuner bushing during this operation. The drill should be run at slow speed if a variable speed unit is available; otherwise, use a Variac to adjust the speed of the drill.

The discussion so far has been concerned with the plate circuit assembly. This assembly can now be temporarily laid aside while the cathode parts are assembled.

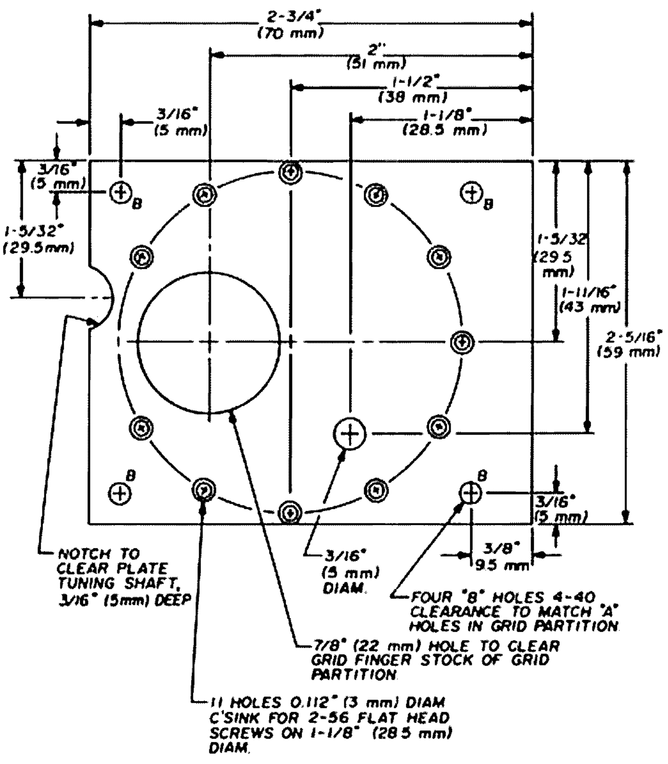
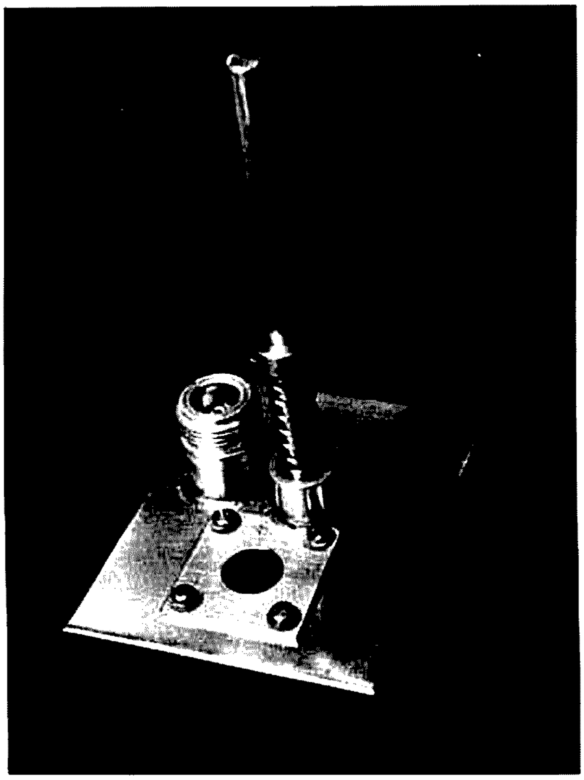


fig. 9. Grid cavity plate. Material is 0.093" (2.5mm) brass.



Cathode partition before the mounting holes have been drilled and the heater/cathode connector is installed.

Fig. 12 is a cut-away view of the cathode assembly showing the cathode plate with its cavity, the type-N input connector and the piston tuning capacitor and tuning sleeve. The type-N connector and the tuner bushing are assembled and soldered in the same manner as described for the output plate. The input coaxial coupler is identical to the output coaxial coupler.

Fig. 10 is an outside view of the cathode partition. There are eleven 2-56 clearance holes located 30° apart on a 2-1/8 inch (54mm) diameter circle, three of which are countersunk to provide a flush surface for the heater/cathode assembly. These holes should not be drilled until later in the assembly.

The four C holes serve a dual purpose. They are primarily screwdriver clearance holes to facilitate the final assembly of the amplifier. However, they can also be used to attach support

Front view of the cathode partition. Notch on rear edge provides clearance for plate tuning piston.

rods for mounting the amplifier to a panel.

cathode heater assembly

Parts from a surplus 2C39 amplifier were used for the heater-cathode assembly shown in fig. 11. The 7/32-inch (5.5mm) clearance hole at the center of the 1-inch (25.5mm) square plate is cut undersize and carefully reamed for a force fit over the heater-cathode assembly. It is important that a 2C39 be plugged into the heater cathode assembly during the reaming process, and also when the plate is soldered in position. Use solder sparingly so that it will not flow between the serrations and onto the 2C39 cathode sleeve. The location of the square plate is critical. Notch one side to clear the tuner bushing.

Before drilling the eleven clearance holes on the cathode partition for the 2-56 mounting screws, assemble the heater/cathode assembly to the cathode partition using shoulder washers and a 10 mil (0.25mm) Teflon insulating sheet. Use the heater-cathode assembly as a template for locating the four mounting holes. Drill and tap these four holes for 2-56 screws. Then attach the cathode cavity to the grid cavity plate (fig. 9) using flat-head, 2-56 stainless-steel screws.

The plate circuit assembly should now be mated with the cathode assembly. Before this can be done a notch must be cut on the grid cavity plate to clear the plate tuning shaft. Next, lay the grid cavity plate face down on the outside of the grid partition and insert the 2C39 into the plate assembly socket. Then plug the cathode partition into the cathode end of the 2C39. If the instructions have been carefully followed, the face of the cathode plate will mate with the cathode cavity ring. If it does not, shim the heater-cathode assembly with a 1-inch (25.5mm) square, thin brass sheet with a central clearance hole and matching mounting holes.

Make sure that the sides of the cathode partition and grid cavity plate

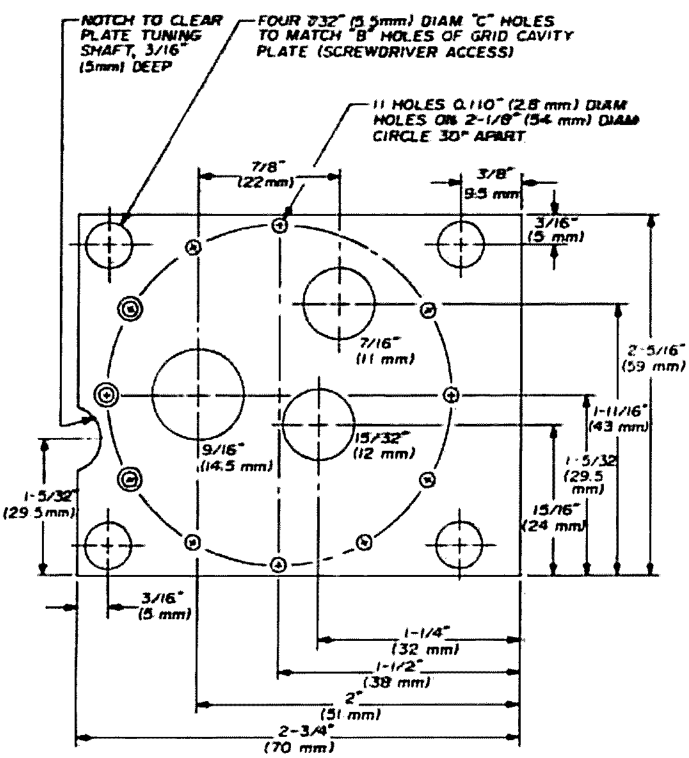


fig. 10. Cathode partition. Holes marked with letter C are screwdriver clearance holes to facilitate assembly. Material is 0.093" (2.5mm) brass.

are parallel and that they, in turn, are parallel with the sides of the plate assembly. Using a scriber, carefully mark the location of the cathode cavity on the cathode plate. This will help to locate the center of the 2-1/8 inch (54mm) diameter circle and the positions of the eleven cavity mounting holes. After these holes have been drilled, reassemble the parts and use the cathode partition as a template to locate the eleven 2-56 tapped holes on the cathode cavity.

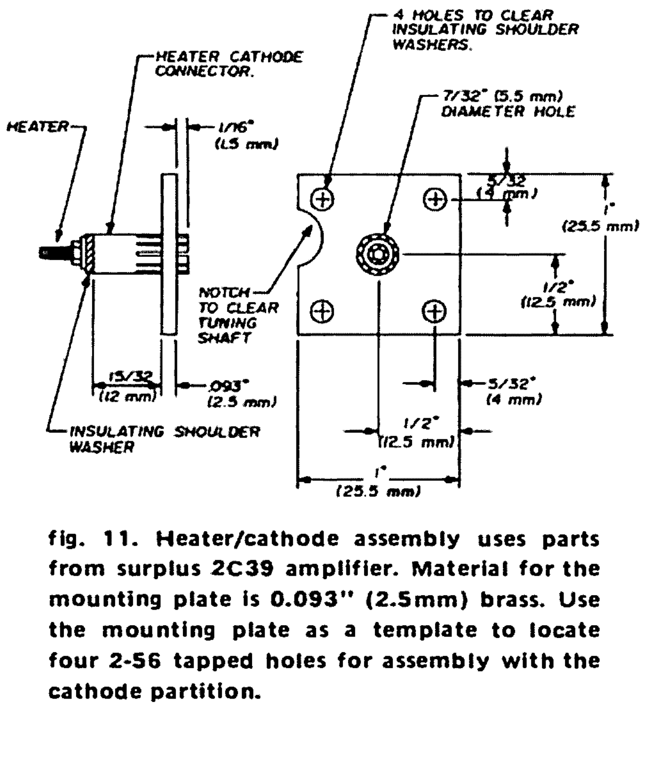
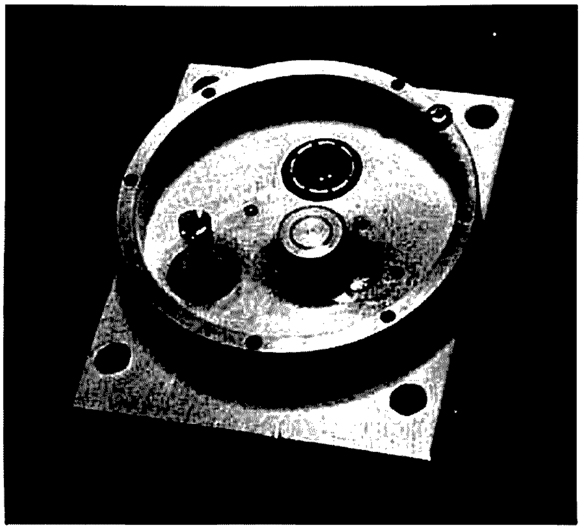


fig. 11. Heater/cathode assembly uses parts from surplus 2C39 amplifier. Material for the mounting plate is 0.093" (2.5mm) brass. Use the mounting plate as a template to locate four 2-56 tapped holes for assembly with the cathode partition.

Insert the 1/4-inch (6.5mm) shaft of the cathode tuning piston through the cathode tuner bushing from the inside of the cavity; then reassemble the cathode partition (fig. 10) with the cathode cavity assembly.

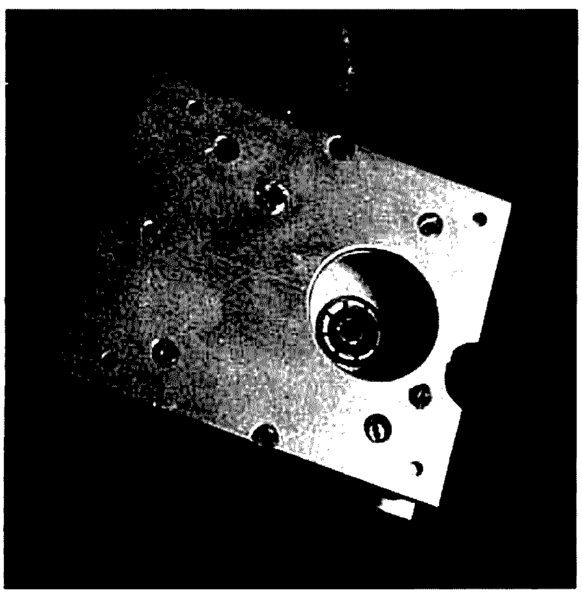
At this point the two major assemblies are complete. It is now only necessary to attach these assemblies together by threading four 4-40 screws through the B holes of the grid cavity plate into the four tapped A holes of the grid partition. A screwdriver can be slipped through the C holes to tighten these screws.



Inside view of the cathode partition and cavity assembly before it is attached to the plate assembly.

tune up

Check the insulation under the heater/cathode plate with an ohmmeter to make sure the filament and cathode are insulated from ground. Also, check the plate socket insulation. Wire the amplifier as shown in fig. 13. Start the tune up by applying a low voltage to the plate, or bring the plate voltage up slowly with the aid of a variable-voltage transformer.



Bottom view of tbe cathode assembly.

The 50-ohm resistor in the cathode return circuit should be adjusted for a quiescent (no-drive) 2C39 plate current of approximately 40 mA with 1000 volts on the plate. With sufficient driving power, the plate current should reach approximately 120 mA. (If a 2C39 equipped with a water jacket is used,[4] the plate current can safely be driven to 200 mA.) The 0-250 mA meter installed in the cathode circuit allows the grid current to be measured as the difference between the cathode and plate meter readings. Grid current levels over 50 mA may be reached.

I recommend the use of a 2304-MHz driver capable of providing about 5.0 watts output to compensate for losses associated with interconnecting cables and fittings. A 2C39 doubler, using a

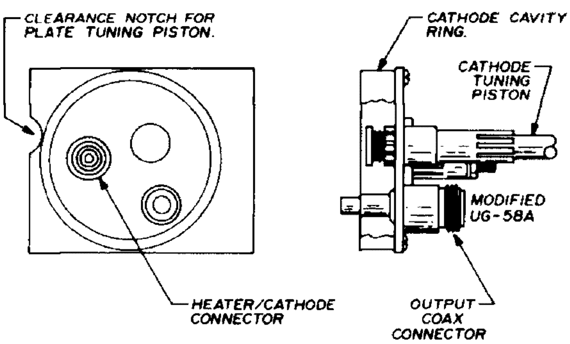


fig. 12. Inside view (left) and cutaway view (right) of the cathode assembly, showing location of the various components.

plate circuit identical to the one described here, will easily provide 5 watts of drive power. With 5 watts of drive the rf power output from the amplifier should be between 20 and 40 watts, depending on the particular 2C39 used in the circuit.*

Surprisingly, the power gain of most 2C39s tried in this circuit measured

*A Hewlett-Packard 434A Colorimetric Power Meter was used in conjunction with a 2- to 4-GHz Narda 10-dB coaxial directional coupler for making power measurements.

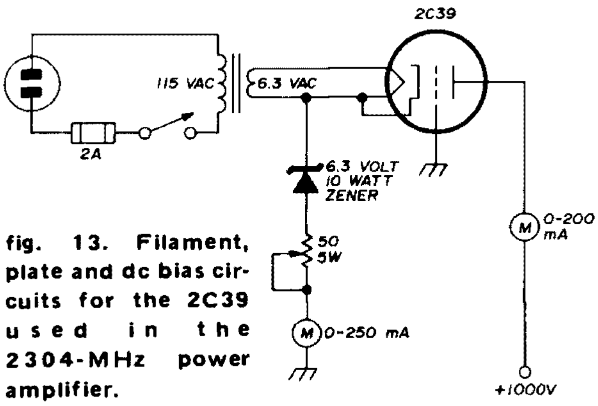


fig. 13. Filament, plate and dc bias circuits for the 2C39 used in the 2304-MHz power amplifier.

about 13 dB. A *hot* 2C39 may deliver over 30 watts CW output with only 1.5 watts drive. Plate circuit efficiencies run between 20 and 25%. Typical operating conditions with air cooling may be 130 watts dc plate power input and 30 watts rf output. The 2C39 plate dissipation would then be 100 watts and the efficiency 25%.

If water cooling is used, over 40 watts rf output can be obtained for 200 watts dc plate input. With air cooling, 15 seconds or more may be required before full power output is achieved once the amplifier has been previously tuned up hot. With water cooling the key-down operating temperature is much less than for air-cooling, and full power output is achieved within a few seconds after plate power is applied.

By following the instructions given in this article, you can generate relatively large amounts of rf power on 2304 MHz. Now, who has a good 2304-MHz eight-tube ring-amplifier design?

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ham radio

bandpass filters

for receiver preselectors

Design and construction details for bandpass filters suitable for many amateur applications

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There was a time when the capabilities of an amateur receiver were well summarized by specifying its selectivity, sensitivity and stability. While the specifications offered by the manufacturers of our present-day receivers rarely include much more, the parameters of significance to a critical operator on the high-frequency bands also include the blocking level, intermodulation levels and sensitivity to cross modulation.

A severe test of an amateur receiver is during contest operation when a large number of signals are present, many of them quite strong. Undoubtedly the most extreme conditions are presented during the ARRL Field Day when the operator must fight not only extensive QRM present on most of the bands, but must also contend with other transmitters operating from his own location, often separated only a few hundred kHz from his own frequency. Designing a receiving system to survive such an environment is one of the most exciting and challenging problems presented to the devoted contest operator.

The solution to large-signal problems lies in the design of the receiver front-end. Great care must be used in determining a proper gain distribution. Further, the proper active devices must

be carefully applied to realize an optimum dynamic range. Along with these requirements, the receiver must be protected from out-of-band signals as much as possible. This latter requirement is met with carefully designed preselection filters and forms the basis for this article. Clearly the design of bandpass filters is applicable to many areas other than receiver preselection.

As with most areas of interest to the technically inclined radio amateur, the preselector synthesis problem can be approached from both a theoretical and an experimental point of view. In this article I will attempt to emphasize the empirical approach. However, any experimental activities are markedly enhanced by an understanding of the basic principles. To this end, some of the fundamentals of filter design will be discussed. Some practical designs are also presented for duplication by the experimenter, if desired.

In a recent article by Nagle,[1] the design of bandpass filters was presented using the classic lowpass to bandpass transformation. While this technique is extremely useful for many design problems, it is generally limited to filters with wider bandwidths. When you attempt to build filters of only a few percent bandwidth, you find that the results are often inconsistent with the classic image-parameter designs. The reason for this is that the experimental results are "distorted" by the finite unloaded Q of the tuned circuits used in the filter. Hence, a more meaningful theoretical approach is to use "predistorted" design tables which allow the designer to account for the unloaded Q

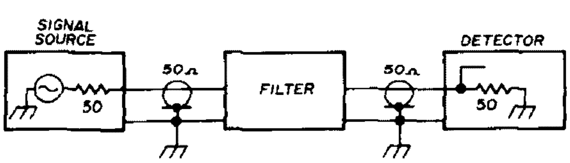


fig. 1. Basic test set-up for aligning and evaluating filters.

of the resonators on hand. The term "resonator" is used in preference to the more usual "tuned circuit" since the methods are applicable to systems at all frequencies from audio to the microwave region.

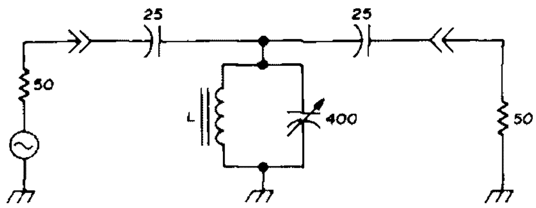


fig. 2. Single resonator filter for operation at 7 MHz. The inductor, L, is 1.3 µH, consisting of 17 turns number-22 wire on an Amidon T-68-6 toroid core; unloaded Q of the inductor is 275. Response of this filter is plotted in fig. 6.

measurement techniques

Shown in **fig. 1** is a generalized block diagram of the test set-up which should be used for the alignment and evaluation of preselector filters. There are three basic parts to the system: a signal source, the filter being tested and a calibrated detector. There are a number of pieces of equipment which could be used for both the signal source and for the detector, depending upon the gear available in your own lab.

One suitable signal source would be one of the many inexpensive signal generators on the market such as the Heath IG-102. However, it is quite important that the output impedance of the source be constant and known, typically 50 ohms. This is rarely the case with inexpensive signal generators. This problem is easily solved by inserting a 10-dB attenuator between the generator and coax leading to the filter. Similarly, a vfo-controlled low-power transmitter covering the frequency range of interest would be suitable if it is used with a 10-dB pad.

Although high-frequency oscillo-

scopes, wide-range spectrum analyzers or even a QRP power meter[2] are all suitable as detectors, probably the most commonly available item is your station receiver. As with the generator, it is quite important that the input impedance of the detector be 50 ohms, rarely the case with receivers. Again, a 10-dB pad at the input to the receiver is a suitable solution. If you have great faith in your receiver's S-meter, you can use it as the output indicator. A much safer method would be to precede the receiver with a step attenuator and

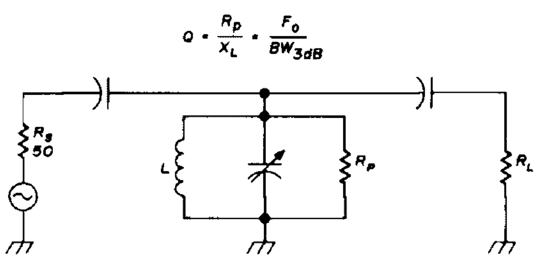


fig. 3. Equivalent circuit for modeling the Q of the inductor used in the circuit of fig. 4 (see text).

monitor the receiver output with an audio voltmeter. Since all measurements will be done with the attenuator using the substitution technique, the audio voltmeter does not need to be calibrated. The attenuators described by Daughters and Alexander[3] are very inexpensive, easy to build, and usable into the vhf spectrum.

The test equipment I use is variable, but typically starts with a homebrew signal generator with about 4-milliwatts output. This output is split, with one component feeding a frequency counter; the other output feeds an attenuation pad which then drives the filter under test. The output of the filter drives a step attenuator which drives a broadband amplifier. This, in turn, is applied to a square-law detector using a hot-carrier diode. The system is suitable for measurements over a range greater than 50 dB at frequencies up to the low end of the vhf spectrum. All measurements presented in this article were obtained with an HP-8640B Signal Generator and a Tektronix 7L13 Spectrum Analyzer, used as a detector. However, all alignment and initial evaluation was done with the less exotic gear available in my home workshop.

the single tuned circuit

As an initial step in our investigation of filters, let's take a look at the common, single resonator. While this configuration is hardly profound, it is typical of the minimal preselection found in most of our amateur receivers. Secondly, many of the conclusions you reach in pursuing such a simple system are qualitatively very general and can be applied in building more elaborate, multi-resonator filters.

For our design example, the resonator shown in **fig. 2** will be considered. The coil is merely 17 turns of number-22 enamelled wire on an Amidon T-68-6 toroid core. This will be resonated at 7 MHz with a 400-pF mica compression trimmer capacitor. The unloaded Q of this inductor was measured as 275. As will be shown, a laboratory Q-meter is not necessary for this measurement. Energy is coupled into and out of this resonator with a pair of 25-pF capacitors. In all analysis, always assume that the filter is driven and terminated by 50-ohm resistive sources.

To analyze this circuit, one more element is needed: Some means for modeling the Q of the inductor. This is shown in **fig. 3**. As is well known, the finite Q of any lumped tuned circuit can be represented by either a series or parallel resistance connected to ideal inductors and capacitors. For our application the parallel representation is more useful and yields R_p = 15700 ohms.

Although I will not go through the details, the resonator is easily analyzed using classic ac circuit theory.[4] First, it can be shown that the 25-pF coupling capacitors have the effect of transform-

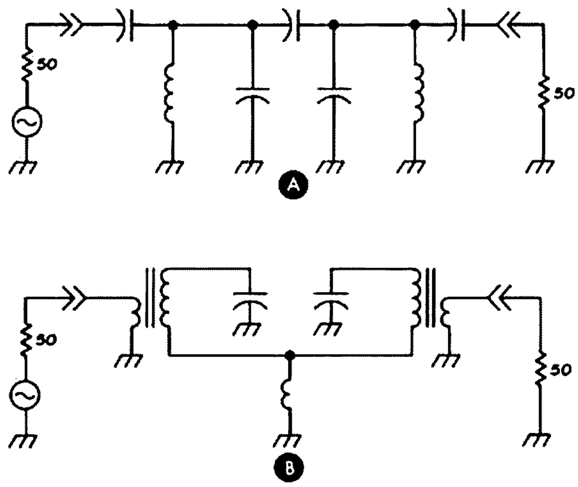


fig. 4. Two forms of the double-tuned resonator. Capacitive coupling is used in (A); inductive coupling is used in (B). Performance of both circuits is identical.

ing the 50-ohm source and load to parallel equivalents. In this case, the series 50-ohm resistor and 25-pF capacitor transforms to a parallel resistance of 16.6 kilohms in parallel with a capacitance just under 25 pF. The net load across the resonator is now the parallel combination of the two 16.6-kilohm external loads and the 15.7-kilohm resistor representing the inductor losses, or, in this case, 5.5 kilohms. Using the equation in fig. 3 which relates Q to parallel load resistance, the loaded Q is calculated to be 95. Additional arithmetic will show that this filter has a bandwidth of 74 kHz and insertion loss of 3.7 dB.

The data on this resonator become more enlightening as you consider some other component values. For example, if you change the input and output coupling capacitors to 10 pF, the loaded Q goes up to 210, yielding a filter with a bandwidth of 33 kHz, but with an insertion loss of 12.5 dB. If you use 50-pF input and output capacitors, you realize a filter with a Q of 32, bandwidth of 220 kHz and insertion loss of only 1.1 dB.

Probably the most significant conclusion is that insertion loss *must* increase as you go to narrower bandwidths. For the single tuned circuit it can be shown that the loss is given by

$$\text{Insertion loss} = -10 \log_{10} \left(1 - \frac{Q_L}{Q_U}\right)^2 \text{ dB}$$

From the equation you see that filters with low insertion loss can be realized only by using resonators with a very high unloaded Q, or by accepting a wider bandwidth (i.e., reduced loaded Q).

While all of this may seem to be

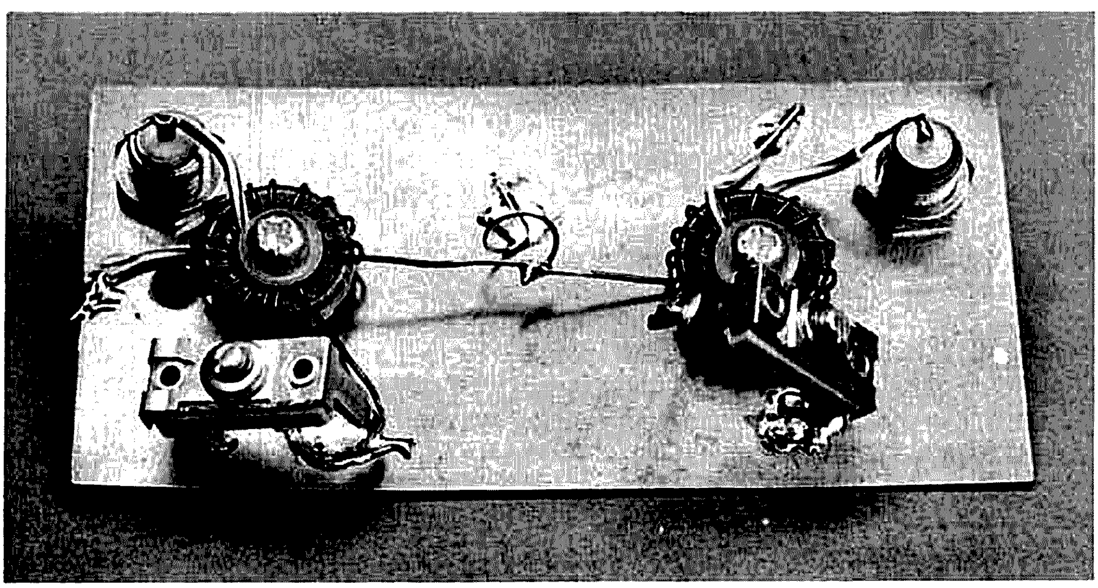


fig. 5. Two-section, double-tuned filter for use on 7 MHz. Measured response of this filter is plotted in fig. 6.

quite basic and perhaps academic, it is quite significant, for it allows you to make measurements which will tell you a lot about your filter. For example, if you measure the bandwidth and the insertion loss of the filter shown in **fig. 2**, you can then calculate the unloaded Q of the resonator. Alternately, if you use a 1-pF capacitor to lightly couple to the load and the source, the insertion loss will be 46 dB, but the measured Q will be within 1% of the unloaded Q of the system. This method of Q measurement is straightforward, and applicable at frequencies well outside the range of the typical Q-meter.

Before progressing to the double-tuned circuit, there are a couple more calculations which are enlightening. If you disconnect the generator but leave the output terminated in 50 ohms, the input resistance seen at the input to the filter is about 102 ohms, and not the 50 ohms you might expect. When working with filters, impedances are not necessarily matched in the classic sense of maximizing power transfer. Similarly, if the input impedance of a receiver is specified as being 50 ohms, this means that the unit should be driven from a 50-ohm source. However, the impedance which would be seen by a bridge looking at the input may be something quite different.

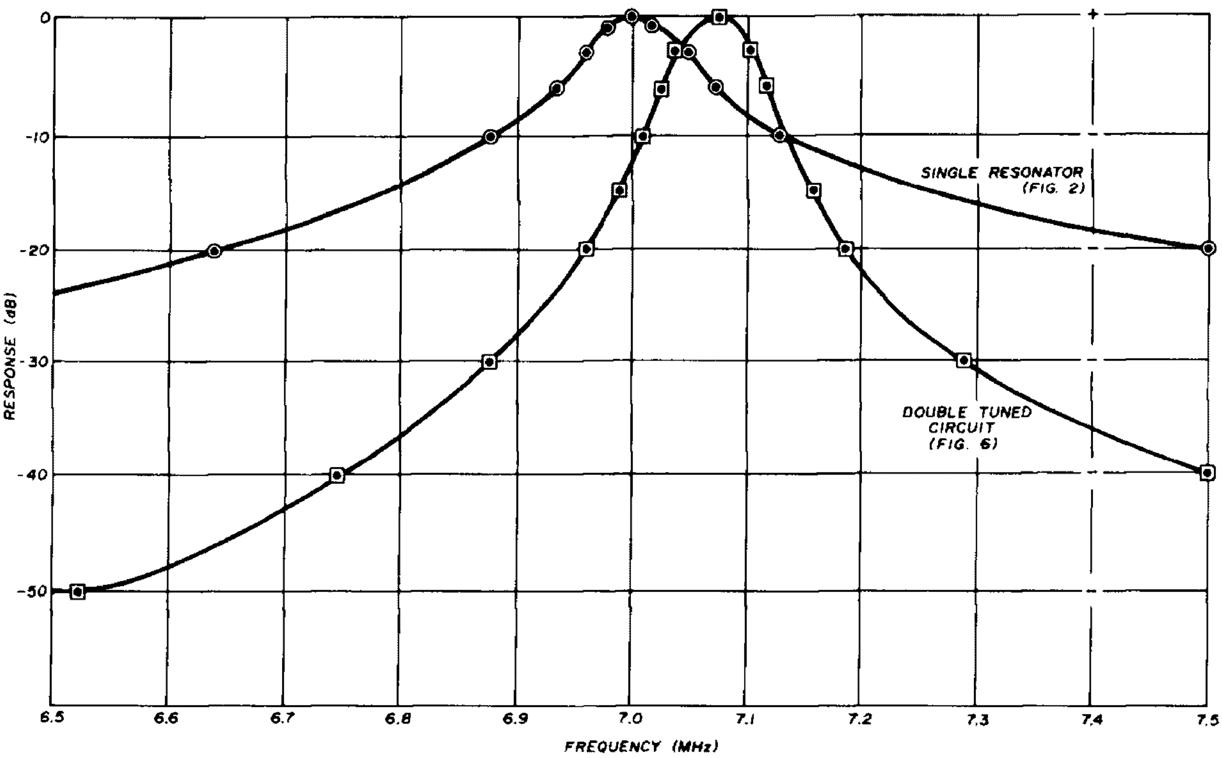


fig. 6. Measured frequency response of the single- and double-tuned 7-MHz filters.

In the example of fig. **2** series capacitors have been used to transform the generator and termination to resistances useful for loading the filter. However, the more common, and often more convenient, method is to use link coupling. Toroid cores have as one of their virtues the asset that impedances transform as the square of the turns ratio. Assume, for example, that a three-turn link were placed on the 17-turn coil and that the link is terminated in 50 ohms. The effective resistance across the coil is thus

$$50\left(\frac{17}{3}\right)^2 = 1606 \text{ ohms}$$

This is, of course, in parallel with any other loads which may be present, including the resistance representing the unloaded Q of the resonator.

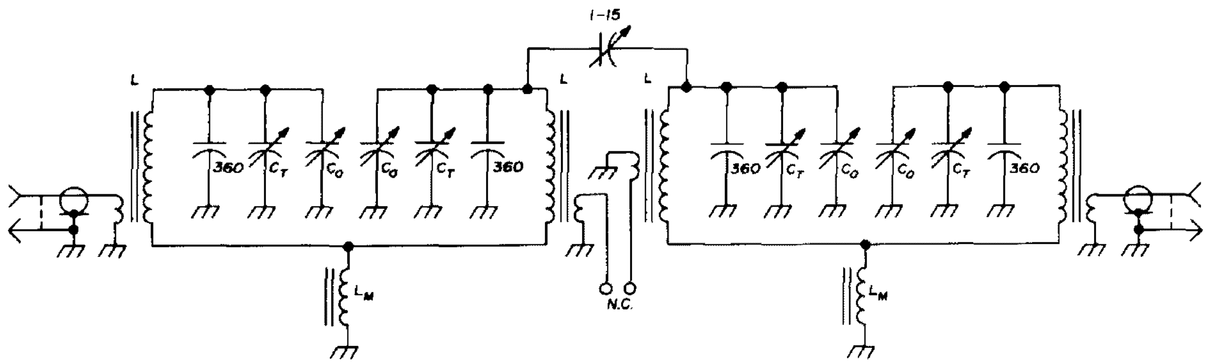
the double-tuned circuit

Shown in **fig. 4** are two forms of the double-tuned circuit. Although a bit more complicated than the single resonator, it has the property that steeper skirts can be realized while maintaining a wider 3-dB bandwidth. As might be expected, you must pay the price of higher insertion loss to realize these assets.

Fig. 4A shows a capacitor for coupling between resonators as well as capacitive coupling to the external load. **Fig. 4B** shows the use of inductive coupling. For fixed-tuned filters the two methods, or mixtures of the techniques such as link loading with capacitive coupling between resonators, are virtually equivalent. However, if the filters are to be tuned over some band of frequencies, a little more care should be taken. If, for example, a dual-section variable capacitor is to be used, the scheme of **fig. 4B** should be used. If inductive tracking is to be used (á la *Collins receivers*), then capacitive coupling should be used between resonators.

Consider an empirical approach to designing a two-resonator filter. The first step is to choose suitable components. Generally, the only criterion of significance is that the unloaded Q be as high as possible. If capacitive coupling between resonators is to be used, it is generally advisable to lean toward lower L to C ratios in the basic resonator since this makes adjustment a little easier. Once the resonators are chosen, each resonator is loaded lightly and equally and the two resonators are coupled lightly. The generator is set at the desired center frequency and the system is tuned to resonance. The insertion loss is measured and then the generator is tuned over the range of interest. A single peak is typically noted.

If the bandwidth is too narrow and/or the insertion loss is higher than acceptable, the coupling between resonators is increased until the passband response begins to appear flat. If the coupling is increased further, a double-humped response will be noted and the insertion loss at the center of the filter will increase. At this point, the loading of the two sections must be increased,



L	33 turns no. 20 enamelled on Amidon T-106-2 toroid cores (approximately 14 μH). Links are each 2 turns
L_m	6 turns no. 20 enamelled on Amidon T-30-2 toroid core
C_O	4-section air variable, 10-160 pF per section
C_T	35-pF trimmer capacitors

fig. 7. Four-resonator filter designed for the amateur 160-meter band. Measured response of this filter is plotted in fig. 9.

the filter re-resonated at the passband center and the coupling adjusted for a fairly flat response. For the class of filters considered in this article it is important to keep the loading equal on each section. This general procedure is continued until the desired bandwidth is obtained, always remembering that there is going to be a tradeoff between bandwidth and insertion loss.

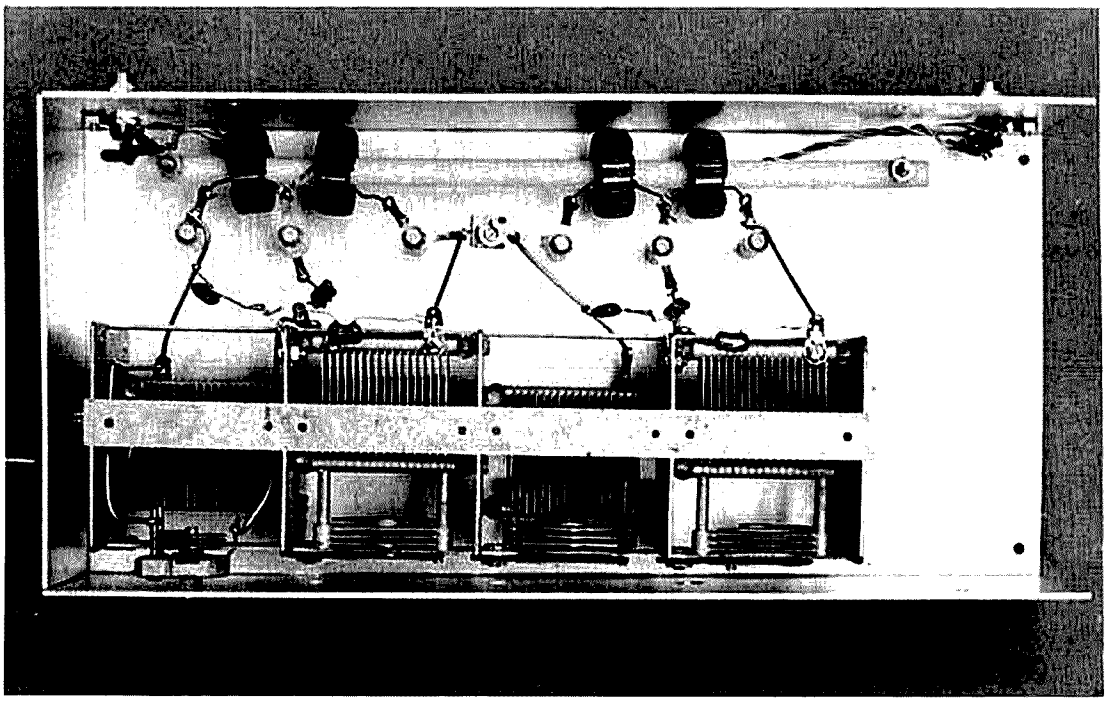


fig. 8. Construction of the four-resonator 160-meter filter shown in fig. 7.

Shown in fig. 5 is a photograph of a two-section, 40-meter filter which was built in this manner. A pair of the 17-turn toroids discussed earlier were used with a 400-pF compression trimmer for tuning. Loading was accomplished with one-turn links. A mutual inductor was used for coupling between resonators. The coupling inductor was one turn, about ¼-inch (6-mm) long and ¼-inch (6-mm) in diameter. Shown in fig. 6 are the measured responses of the single-resonator (fig. 2) and double-resonator filters. The superior skirt response of the two-resonator system is obvious. The insertion loss of the double-tuned system was measured at 7.5 dB, an acceptable figure for typical 40-meter work. However, this much insertion loss would be clearly intolerable on the vhf bands, or even on 10 meters.

There is a slightly more formal, but extremely useful, method for adjusting a two-section, equally-loaded filter. In the earlier discussion of the single-resonator filter it was noted for the 7-MHz filter that very light loading occurs as you couple into the system with 1-pF capacitors. This was used to advantage in measuring unloaded Q. This light "probing" of a resonator is also used in this filter tuning technique.

In a two-section filter designate the resonator driven by the generator as resonator A; resonator B is connected to the load. The two resonators are assumed to be identical. The following tuning procedure is followed:

1. Resonator B is shorted at the hot end. The generator and detector are each lightly coupled (i.e., 1-pF capacitors) in an identical fashion to resonator

A. This resonator is tuned for resonance at the center frequency and the couplings to the generator and load are checked for at least 30-dB of insertion loss.

2. The ultimate design bandwidth is picked and designated as BW. The coupling element between resonators A and B is arbitrarily adjusted to some level and resonator A is re-trimmed for a peak response.

3. While still lightly exciting at the center frequency and probing resonator A, resonator B is unshorted and tuned to resonance. This is detected as a large dip in output in the detector, often as much as 30 dB in tightly coupled, high-Q resonators.

4. Now, the generator is detuned from the center frequency of the filter. As the generator is swept through the range of interest, two strong peaks will appear in resonator A. These frequencies are noted.

5. Steps 2 through 4 are repeated with different adjustment of the coupling between resonators A and B. The coupling is correct when the frequency separation between the peaks (step 4) is BW x 0.707.

6. Now, resonator B is again temporarily shorted. The detector is left coupled very lightly to resonator A. However, the generator is tightly coupled to the 50-ohm generator. Resonator A is tuned for a peak response at the center frequency and the loaded 3-dB bandwidth is measured by sweeping the generator.

7. Step 6 is repeated while adjusting the loading until the loaded bandwidth of resonator A equals 0.707 BW.

8. The detector is now lightly coupled to resonator B and resonator A is shorted. Resonator B is tightly coupled to the generator and the loading adjusted as in steps 6 and 7 for a loaded bandwidth of 0.707 BW.

The filter is now tuned. While this method sounds a bit cumbersome, it's really much easier to do than it is to describe. The resulting filter will have something close to a two-pole Butterworth response. As a general rule-of-thumb, the insertion loss will be around

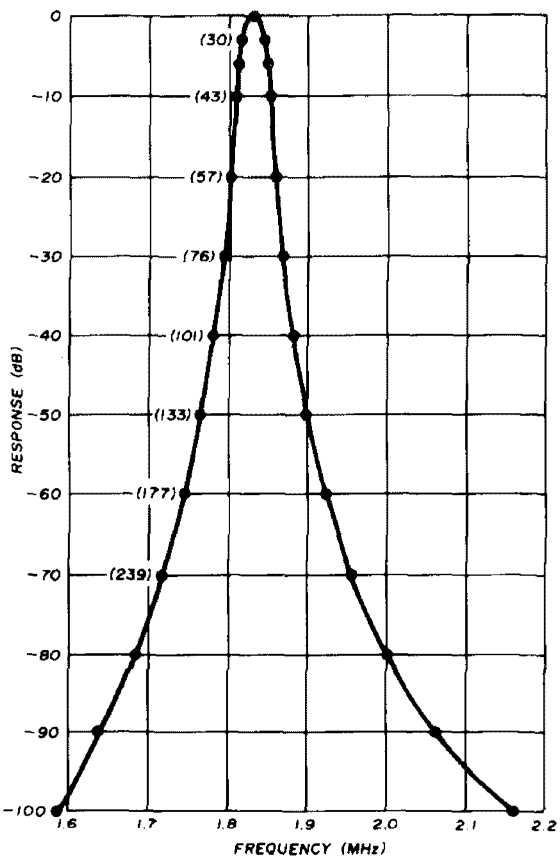


fig. 9. Measured frequency response of the four-resonator 160-meter filter shown schematically in fig. 7.

4 dB if the design bandwidth is four times the unloaded bandwidth of the resonators chosen. Decreasing the bandwidth to just twice the unloaded resonator bandwidth will increase the insertion loss to around 10 dB. If the coupling between resonators is made a bit tighter and the ends of the filter are loaded lighter, a Chebyshev response results. This will steepen the skirt response at the expense of additional insertion loss.

four-section design examples

As you progress to filters with more

than two sections, the problems also increase. As expected, the insertion loss increases as the number of resonators goes up. However, the skirt response also becomes better. In many receiver applications, the noise figure degradation resulting from a higher insertion loss makes one shy away from filters which are overly exotic. A better approach is often to place a low-gain amplifier between a pair of double-tuned circuits.[5]

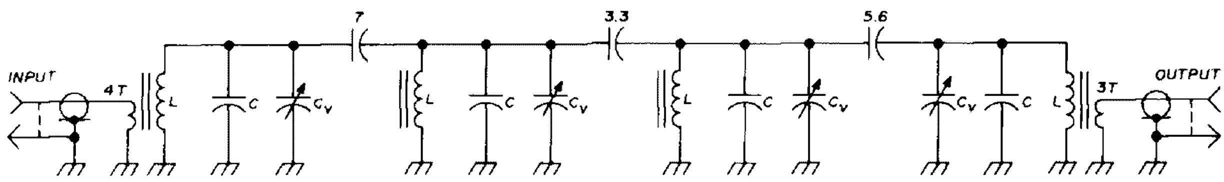
One empirical approach to the design of a four-pole filter is illustrated in the circuit of **fig. 7**. This unit was built for use at W7RM on the 160-meter band. The filter is tunable from 1.8 to 2.0 MHz and has an insertion loss of 5 dB. The 3-dB bandwidth is 30 kHz and the 6- to 60-dB shape factor is 4.78. Stop-band attenuation is more than 120 dB.

The key to the performance of this filter is the high-Q toroid cores chosen for the resonators. Amidon T-106-2 toroids were used since they exhibit an unloaded Q of 330 at 1.8 MHz, a feat difficult to achieve with air-core coils of any reasonable size. Incidentally, I have found the Q and inductance data supplied by Amidon* to be quite reliable and repeatable.

The filter was built by first constructing two identical double-tuned circuits. They were each adjusted for about 2-dB insertion loss with 50-ohm terminations at each end. Then the two filters were capacitively coupled through a small air trimmer. Final adjustment was done by trimming the frequency of the resonators and the coupling capacitor for optimum pass-band shape. The fact that one capacitive coupling element was used in a system with capacitive tracking led to a 1-dB variation in insertion loss as the filter is tuned over the band. Also, it was found necessary to adjust the turns spacing on some of the coils in order to implement tracking.

The Top-Band filter is shown in the photograph, **fig. 8**. The resistors near the BNC coax connectors are 3-dB pads which were inserted to insure that the filter termination is always fairly close to being correct. This is often a problem on the 160-meter band where split frequency operation is used. Shown in **fig. 9** is the measured response of the filter when tuned to the European 160-meter "window" at 1830 kHz.

Presented in **figs. 10** and **11** is data for a filter for the 80-meter band. This filter was designed using the pre-distorted design data presented in Zverev,[6] assuming an unloaded resonator Q of 225. The measured insertion loss of 4.4 dB was slightly under that predicted, indicating that the unloaded Qs were a bit higher. Otherwise, the results are very close to the desired Butterworth design. Although the filter was designed and aligned at 3750-kHz, re-alignment at 3600 and 3900 yielded similar



L	35 turns no. 22 enamelled on Amidon T-68-2 toroid cores (7 μH). Input link is 4 turns, output link is 3 turns	C	210 pF, 1% silver mica
		C_V	60-pF mica compression trimmers

fig. 10. Four resonator filter for use on 80 meters has 100-kHz bandwidth, 4.4-dB insertion loss and 6- to 60-dB shape factor of 5.16. Response is plotted in fig. 11.

*Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607.

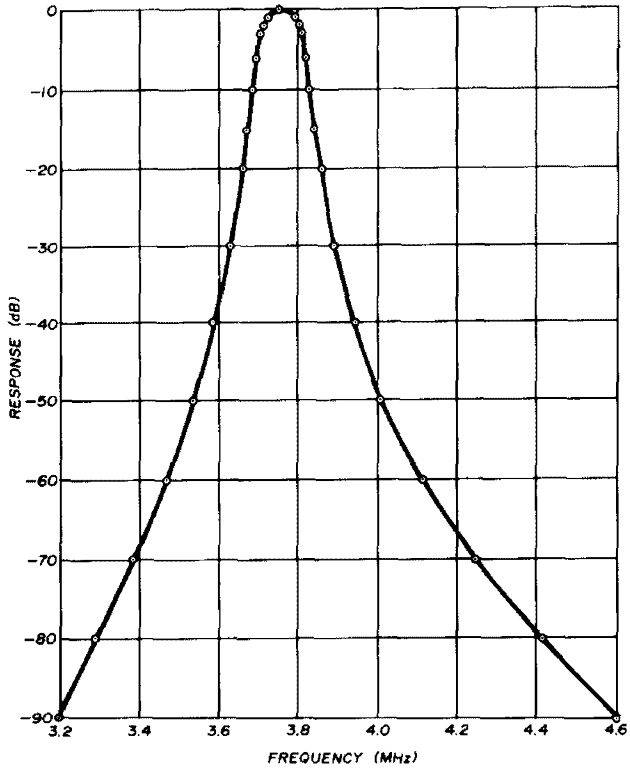


fig. 11. Measured frequency response of the four-resonator 80-meter filter shown in fig. 10.

results. To fit the design exactly as the center frequency is changed, the coupling capacitors between the resonators should be directly proportional to the total tuning capacitance in each resonator.

The synthesis of multipole filters of narrower bandwidths or at higher frequencies becomes more difficult. The four-pole 80-meter filter described above had a bandwidth of about six times the unloaded bandwidth of the resonators, consistent with an insertion loss of a little over 4 dB. If a sharper filter is required, the only solutions are to accept much higher insertion losses, or to use resonators with much higher unloaded Q.

Myers and Greene[7] have reported on the construction of helical resonators* for the high-frequency region. For example, a two-section filter was described using resonators capable of yielding an unloaded Q of 700 at 7 MHz. With such resonators a double-tuned circuit at 40 meters would yield a 4-dB insertion loss with a bandwidth of 40 kHz. When tuned in the CW segment of the band there should be more than 30-dB attenuation over most of the 40-meter phone band. If proper input-output isolation is maintained in such a filter, the attenuation to even-order, harmonically-related bands should be well over 60 dB. While such performance will do wonders for the typical Field-Day installation, proper filter adjustment is mandatory.

Oddly, the basic problems become a little less difficult as you move into the vhf region for it is easier to build high-Q resonators. The procedures are essentially the same as those described, although the systems may look much different physically. The single, most important concept to remember at any frequency is that the price of narrow bandwidth must be paid in insertion loss.

*A helical resonator has been described in the amateur literature as a coil surrounded by a shield. A more accurate description is that the helical resonator is a quarter-wavelength of special helical transmission line. This line is similar to coax except that the inner conductor is helically wound, yielding a very low axial propagation velocity. Hence, an electrical quarter-wavelength will be much shorter than an equivalent coaxial resonator.

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ham radio

principles of speech processing

A discussion of the various speech processing systems, and some suggestions for further experimentation

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Although there is a great deal of interest at present in developing practical speech processing systems to increase the effectiveness of radio transmitters, the basic principles and problems have been understood since the 1920s.[1,2] First of all, most speech falls within a frequency band of a little over three octaves around the range from 300 to 3,000 Hz, and most of the speech energy is concentrated in the lower octave of the range. These lower-frequency sounds contribute to the individual timbre of the voice but have little to do with the intelligibility of speech. Furthermore, most speech intelligence is carried in the band around 1 kHz in most voices. Most ears are most sensitive around this band, too.

Speech has a high ratio of peak-to-average energy. The peaks may be clipped until the speech envelope approaches a series of square waves at clipping levels around 30 dB, where speech is still readily intelligible although unnatural sounding. Due to the wide range of individual voice characteristics, languages and dialects used there is no agreed index of intelligibility,[2] and optimal speech processing systems must be tailored to the individual voice.

Practical speech processors are designed to take advantage of all or some of these facts. In communications practice the designer is prepared to sacrifice voice fidelity for increased communication effectiveness, but it should be kept in mind that very subtle processing techniques are part of the normal practice in recording and broadcasting studios where the object is to keep maximum fidelity of sound within the limits of the transmission medium. These systems are well worth study by anybody developing communications speech processors.

frequency shaping

Since most of the speech energy is concentrated in the lower part of the voice frequency range and contributes little to the transmitted intelligence, it follows that an immediate increase in the transmitter effectiveness will result by ensuring that the transmitter speech amplifier has a falling bass response. The higher-frequency speech components above 2.5 kHz contain little energy or intelligence and are normally restricted in communications systems to reduce the transmission bandwidth. In ssb transmitters the extreme high and low

voice components are usually attenuated by the sideband filter. Moreover, communications microphones are generally designed to emphasize the essential mid-range voice frequencies. Nevertheless, some advantage can be expected by designing the speech amplifier so that it has a frequency response similar to that shown in **fig. 1** where the slope is 12-dB per octave below the knee point at 1.1 kHz.[3]

Good results are also reported by Schmitzer[4,5] using a speech processor with a preamplifier having a passband resembling fig. 1 with a slope of 6-dB per octave below 2 kHz, increasing to 12-dB per octave below 300 Hz.

It is impossible to predict the increase in effective communication power due to frequency shaping alone since individual voices vary, but it is safe to say that the advantage will increase in proportion to the deepness of the voice and the bass response of the microphone. Subsidiary advantages include a reduction in ac supply hum and other low-frequency noise which become a severe problem when high levels of compression or clipping are used following the microphone amplifier.

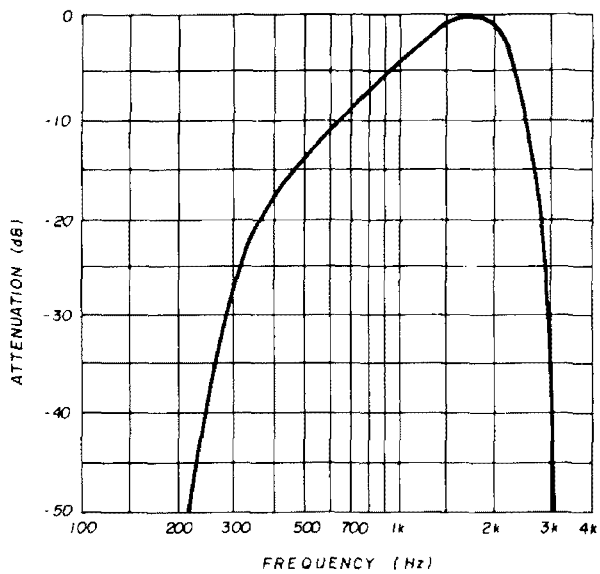

fig. 1. Frequency response of a speech processor with 12-dB per octave rolloff below 1.4 kHz and sharp rolloff above 2 kHz.[3]

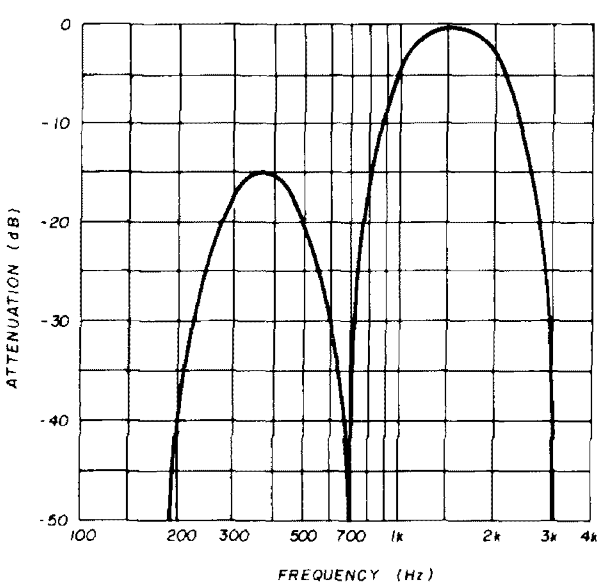

fig. 2. Frequency response of an amplifier that places a notch between the lower and middle speech bands. Theoretically this should give improved intelligibility, especially if the two channels are independently clipped and then summed together.

Listening to transmissions on the air leads one to believe that many signals would have improved intelligibility if the size of the coupling capacitors in the speech amplifiers were reduced, a simple and inexpensive modification.

There is room for experiment in even more radical shaping of the audio bandpass. For example, it's possible that the introduction of a slot into the audio bandpass in the region of 700 Hz, so as to split the band into two parts as shown in **fig. 2**, would permit a gain in intelligibility for a given power. By juggling the output of the upper and lower channels the best trade between intelligibility and naturalness for a particular voice and style of operation should be possible. How wide or deep the slot should be made (or if it should be made at all) is a matter for urgent experiment.

Frequency shaping alone will give advantage in any form of voice transmission, but speech pre-emphasis becomes imperative if any form of compression or amplitude limiting (clipping) is used in the system. In fact, any device

of this type that does not use pre-emphasis is failing to make the best use of its possibilities and demonstrates a lack of understanding of the principles outlined in the first paragraph of this article.

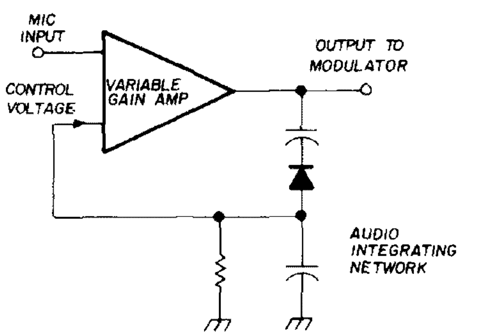

fig. 3. Block diagram of a simple audio compressor.

dynamic compression

Since the amplitude of the energy peaks in speech are much higher than the average energy level, and we know that reducing this dynamic range has little effect on intelligibility within wide limits, it follows that some increase in effective communications power should be possible by applying automatic gain control (agc) to the audio amplifier section of your ssb transmitter. All dynamic compressors work by feeding back a control voltage from some later stage of the system so that gain is varied in such a way that the output level is held more or less constant.

The automatic level control (alc) used in many ssb transmitters functions as a dynamic compressor. Many of these systems work by taking a small control voltage from the grid circuit of AB1 tetrodes which is developed as the tubes run into grid current. Thus, as the tubes are pushed into AB2, the speech envelope of the ssb signal is rectified and fed back to an earlier amplifier, reducing its gain, tending to compress the transmitter output. Many amateurs run their transmitters well into alc and thus have a certain level of compression without realizing it. When additional speech processing is added to the system the result may not be as good as expected since the system is already providing some compression.

Running such transmitters into alc is a questionable practice. No figures on intermodulation products emitted by equipment as it runs into alc seem to be available. The distortion level on the verge of AB2 operation need not necessarily be high so long as certain design considerations are met: notably low impedance and good regulation in the grid bias system. But most alc systems run by virtue of having a considerable impedance in the grid circuit over which the alc voltage is developed and there is likely to be a sudden rise in splatter products as alc operates.

Audio compressors acting on the agc principle have been critically discussed elsewhere[5,6] and many practical designs have been described.[7,8,9] With sophisticated designs intelligibility in noise may be increased up to about 4 dB. The chief disadvantage of simple compressors is that in the intervals between words the gain rises and, thus, background noise appears to rise. Tailoring the time constants in the agc loop cannot give a very high modulation index: if the time constant is fast enough to follow the fast speech sounds then the system is pushed into clipping and heavy distortion can occur. A slow time constant means that initial sounds overmodulate before the system can compensate for them.

Compressors, especially those without careful audio-frequency pre-emphasis, may actually degrade the intelligibility of the transmitted signal while appearing to put out more power since a receiver S-meter may read considerably higher due to the fact that the compressor is integrating background noise into the transmission.

audio clipping

Like an audio compressor, an audio

amplitude limiter is an apparently simple device that may be inserted in series with the transmitter microphone input. If used intelligently it is possible to obtain at least 6-dB increase in effective power, equivalent to an input power increase of four times, albeit with considerable loss in the natural quality of the voice.[10] Properly adjusted audio clippers prevent overdriving later transmitting stages.

All clippers operate by setting up a stage that will pass signals up to a certain amplitude but limit all signals greater than this level. The net effect of this is to put a flat top on the speech envelope which, at extreme clipping levels, approaches a train of square waves. Fourier analysis and practical experience show that a square wave with a 1:1 duty cycle generates the fundamental plus odd harmonics. If the square wave becomes even slightly asymmetrical even harmonics appear, so it follows that care should be taken to ensure that the clipping action is truly symmetrical.

Many published designs are open to the criticism that the clipping is accomplished by a simple pair of diodes. For the system to show its full capabilities some care must be given to the design of the clipper. Carefully matched silicon diodes with forward bias may be adequate, but a better approach is to use a true differential amplifier clipper as shown in **fig. 4**. Another approach is

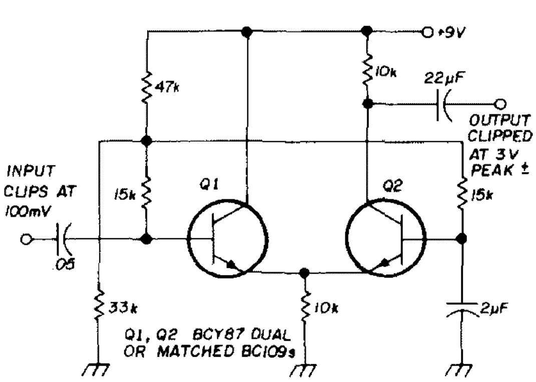

fig. 4. Differential amplifier clipper that provides gain as well as precise and symmetrical clipping.

to use an *operational amplifier-clipper* as shown in **fig. 5**. These have the additional advantage that they produce gain as well as limiting.

The high-order products produced by audio clipping can cause splatter. The

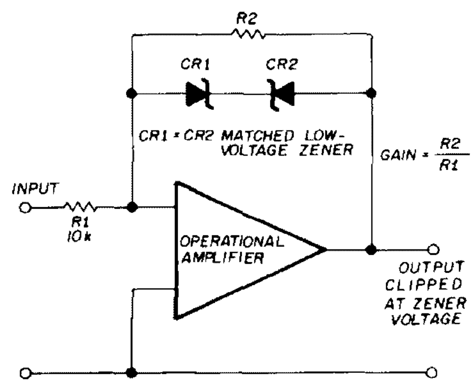

fig. 5. Operational amplifier used as an active audio clipper.

low-order products fall within the speech passband itself and cause distortion of the speech. This is one of the limiting conditions of audio clipping since it is clear that a point will be reached where increased clipping generates sufficient distortion products for intellibigility to drop. It is for this reason that frequency response shaping becomes so important in this system: if low-frequency speech components are reduced by a preamplifier with a bandpass like that shown in **fig. 1**, higher clipping levels can be obtained for a given degree of distortion.

split channel audio clipping

Another approach to the distortion problem in audio speech clippers is to divide the speech spectrum into bands which are independently clipped and filtered. The general idea is shown in the block diagram of **fig. 6**. Although I have not been able to trace a detailed discussion of this system, one reference[14] gives a performance curve of such a device showing results comparable to an rf clipper up to 30-dB clipping level. Another device of this type is under

development by R. Newsome[11] but full results are not yet available.

Another advantage of splitting the speech into channels in this way is that by adjusting the gain of each channel the system can be easily optimized for individual voice characteristics. the disadvantage of this system is the increased complexity. Nevertheless, it is composed of inexpensive components and is an obvious project for amateur experimenters since it can be set up with fairly elementary test gear.

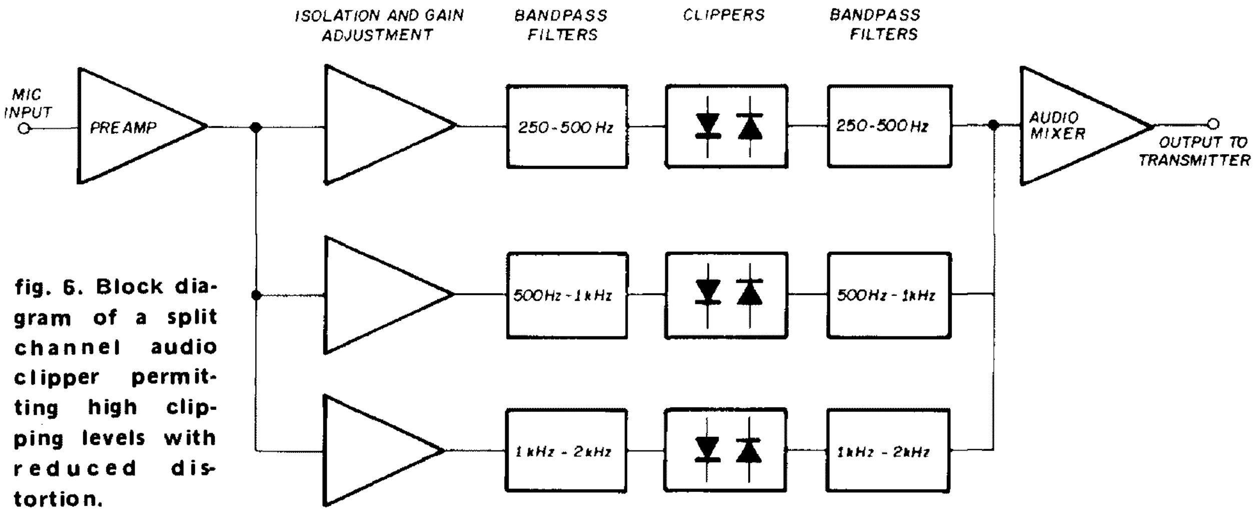

fig. 6. Block diagram of a split channel audio clipper permitting high clipping levels with reduced distortion.

At first glance it seems a simple job to get rid of the high-order clipping products that fall above the speech band. Note that it is also necessary to cut off the upper speech frequencies prior to the clipper; otherwise they would interact with the high order harmonics and produce further intermodulation products that fall in the desired passband. The problem is to produce a filter that sharply attenuates frequencies above about 2.5 kHz. Such a steep filter requires either high-Q LC circuits or active RC filters. Filters of this type tend to overshoot (ring), and in this application this means that when each clipped wavefront passes through the filter, spurious pulses are generated which overload the subsequent stages.

It is clear that the lowpass filter following the clipper must be designed with considerable sophistication if satisfactory operation is to be achieved. This problem has been discussed elsewhere.[4,5] One solution is to use a passive filter section preceding the active filter. By careful choice of values it is possible to achieve an attenuation of around 20 dB per octave above 3 kHz with little overshoot. Such filters should be checked for ringing by feeding 1-kHz square waves through them.

In summary, it can be said that audio clipper-filter type processors can be made to give a very good account of themselves, but considerable care must go into their design. A good example will be as intelligible as an rf clipper up to about 20-dB of clipping although the rf clipper will sound much nicer. A bad audio clipper will sound awful long before 20-dB clipping is reached and, like some compressors, may result in a bigger but less readable signal. Compared with a full rf clipper the single-channel audio clipper shows advantages of cost-effectiveness and may be easily transferred from one rig to another. Audio clippers work well on fm, too.

rf clippers

Since rf clippers have been discussed at length elsewhere,[10,12,13] they will not be discussed in detail here. It is worth noting, however, that few of the discus-

sions on rf clipping mention audio pre-emphasis as described in this article. If audio-frequency shaping is added to an rf clipping system increased effectiveness will result.

It is remarkable how little work has been done on the most elementary form of rf clipping: clipping the double-sideband signal after the balanced modulator and prior to the existing ssb filter. A practical circuit using this system is given in **fig. 7** and has been a great success at ZL1BN; a Yaesu FT-200 so fitted ran up nearly a million points for a national win in the 1971 ARRL DX test, not to mention wins and places in other contests. With the microphone gain up full, the output reads over twice normal without running into alc while voice quality is excellent. Note that the coupling capacitors in the speech amplifiers have been reduced.

This unit represents the greatest increase in transmitter effectiveness per unit cost I have ever seen. Certainly it doesn't have the potential of formal ssb clipping, but heavier clipping of any sort would probably show up the limitations of the power-handling capability of the sweep-tube finals. As things stand they go on year after year without replacement.

The system shown in **fig. 7** is about as primitive as one can get and has the undesirable feature that as the diodes go into clipping they load both the filter and the balanced modulator. This must add distortion to the signal. A more sophisticated system would be to use isolating amplifiers between the modulator, clipper and filter with a proper differential amplifier clipper. Fully developed, the results should be at least as good as the best possible audio processor.

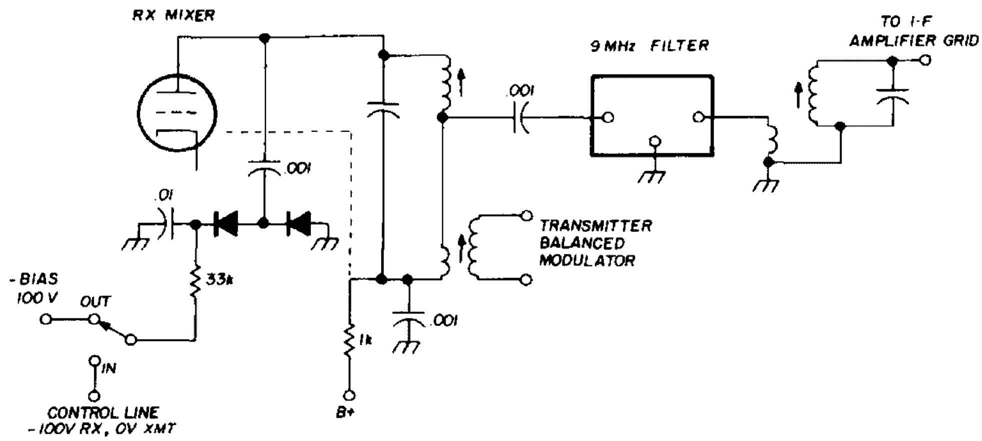

fig. 7. Simple diode clipper operating upon the double-sideband suppressed carrier signal prior to the filter in a Yaesu FT-200 transceiver. Diodes are unbranded high-speed computer types.

It is worth remembering that whatever system is used, extraneous noise becomes more and more prominent in the transmitted signal as clipping levels rise beyond 20 dB or so. It is possible to reduce environmental noise to a degree but it is quite a trick to talk without breathing.* This is a consideration to be taken into account before scrapping a system with moderate clipping for one that might provide more. For practical purposes 30-dB clipping can be taken as an upper limit.

It is generally recognized that the final amplifying chain of the transmitter must have sufficient power capability to take the increased duty cycle of clipped speech ssb. Apart from the risk of thermal breakdown of components,

*It would be good practice to precede the clipping stage with a squelch circuit to eliminate background noise between speech syllables.

operators should check the dynamic characteristics of the final amplifier stage by running a series of dots and dashes from an electronic key and examining the envelope of the CW output with an oscilloscope. Waves in the top of the pulses are usually a symptom of poor dynamic regulation in the power supply and imply a rapid increase in spurious output from the transmitter if run at full bore on clipped ssb.

other systems

In the speech-processing circuit used in the commercial Comdel processor[15] the signal is processed at a low rf frequency and then converted back to audio and fed into the microphone jack of an unmodified transmitter (as in fig. 8). This system has most of the virtues of both rf and audio clipping and it is surprising that amateur versions are not common in the literature.[16] Since the cost of ceramic filters for 455 kHz is dropping relative to the cost of many other components this system would compare very favorably with the more elaborate audio speech processors in cost-effectiveness. This would seem to be the method of choice for the experimenter wanting to increase communication power without making internal changes to his ssb transmitter.

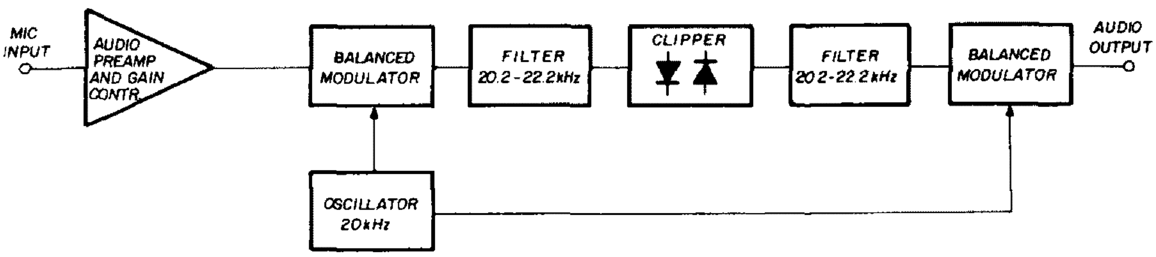

fig. 8. Block diagram of a speech processor in which the clipping is done at a low radio frequency and then converted back to audio to be fed to the microphone input of a transmitter.

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ham radio

phase-locked loop RTTY terminal unit

Easy to build phase-locked loop RTTY terminal unit uses only three ICs and requires no tuned circuits

Nat Stinnette, W4AYV, Post Office Box 1043, Tavares, Florida 32778

Here's a simple RTTY terminal unit which should interest both the beginner and the old timer. It is easy to build and adjust, and does a good job of copying both wide and narrow shift. The circuit, shown in **fig. 1**, uses a 741 op-amp IC, U1, as a limiter, an NE565 phase-locked loop, U2, another 741, U3, as a voltage comparator or slicer, and an MJE340 keying transistor.

This terminal unit requires no filters because it works on the fm principle. Although the theory of the phase-locked loop has been covered extensively elsewhere, a short discussion will be helpful. Simply stated, the incoming signal locks onto a voltage-controlled oscillator, the oscillator frequency is placed between the frequencies of the mark/space tones. As these tones alternate, the output of the PLL can be made to produce plus and minus voltages by connecting a voltage comparator to the output of the NE565. This plus and minus voltage corresponds directly to the mark/space tones and can be used to key the loop circuit of your teleprinter.

This method of RTTY demodulation has the advantage of requiring no tuned filters and will tolerate considerable drift and copy shifts from 170 to 850 Hz simply by properly tuning in the signal.

Construction of this circuit is not critical and a perf board will serve quite well.* A regulated power supply providing ±10-12 Vdc is required. A tuning meter is required and the simplest is a zero-center milliammeter with suitable dropping resistor as shown in **fig. 1**. It indicates a plus current on *mark* and a minus on *space*. When receiving a properly tuned FSK signal the needle tends to hover around the zero center. If a zero-center milliammeter is not available, use a vtvm set to about 25 Vdc and advance the needle to center scale.

*Just prior to publication the author advised that he has designed a PC board for this PLL terminal unit. Undrilled circuit boards are available from his for $4.75 each. Completely wired and tested units are available for $25.95. Editor

alignment

Adjustment is easy: after connecting the loop voltage and the power supply to the terminal unit, check the plus and minus voltages. These can be from 10 to 12 volts but should be within 0.5 volt. The input can be 500 ohms or the speaker output from your receiver. Close switch S1, tune in a good, steady narrow-shift RTTY signal, preferably running a tape. The received tones should be in the vicinity of 1500 Hz. This can be done by ear and is not at all critical. Set potentiometer R1 at the end of rotation (maximum resistance). The zero-center meter should read approximately zero or a little on the plus side. Now advance R1. The meter should move more to the plus side and flicker back toward zero. If it does not, change sidebands if you are receiving in the ssb mode.

If receiving in the CW mode, move the bfo to the other side. Open switch S1 and turn on the teleprinter. As R1 is further advanced a point will be reached where the machine will begin to print and the needle of the meter will stay pretty near zero during copy and maximum plus on mark/hold tone. This is the proper adjustment for R1. As R1 is advanced further the meter needle will move toward the minus side and the machine *will* eventually run open.

With a little tuning practice and watching the meter a signal can be tuned in with no difficulty. After a while you may want to adjust R1 slightly one way or the other to receive signals better at some slightly different tone. However, once it has been properly set it need never be changed. With the audio input off the vco can be heard by connecting headphones between pins 4-5 of U2 and ground.

It is possible to receive both wide and narrow-shift RTTY with one setting of R1, but if the circuit is adjusted to receive narrow-shift, wide-shift RTTY signals are sometimes hard to tune with the same setting. However, since narrow-shift is generally used now, one setting for it should suffice. If you expect to regularly copy both shifts, it would be a good idea to install a second potentiometer with a switch to select either wide or narrow shift.

Switch S1 should be closed while tuning as random receiver noises and other stations will produce garble. It *must* also be closed when transmitting.

ham radio

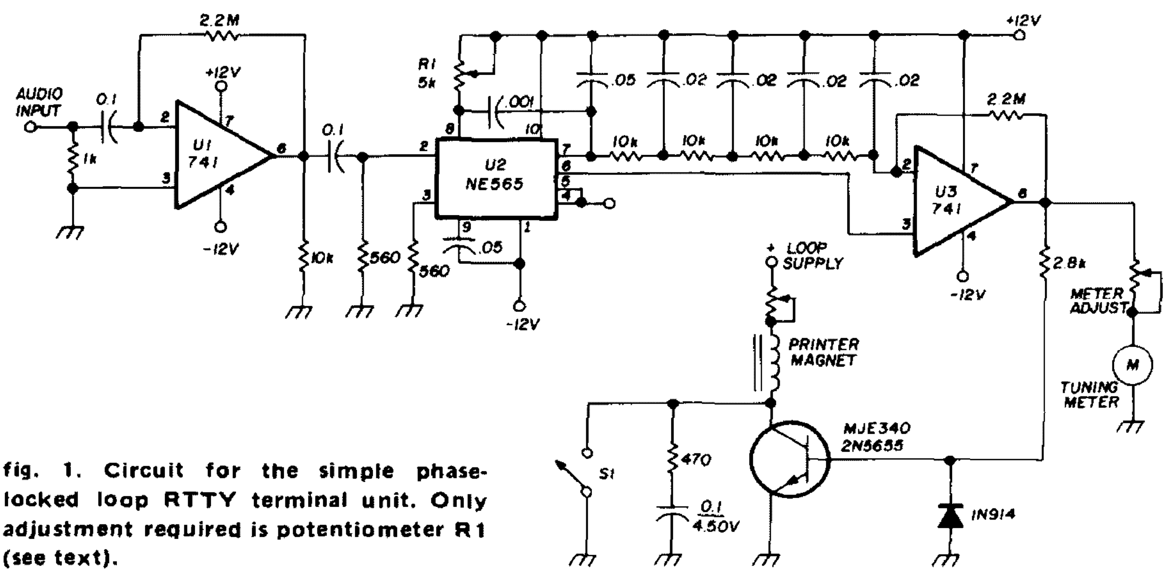

fig. 1. Circuit for the simple phase-locked loop RTTY terminal unit. Only adjustment required is potentiometer R1 (see text).

1200-MHz frequency scalers

New Fairchild sub-nanosecond logic circuits can be used to build frequency scalers that operate to beyond 1000 MHz

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Fairchild Semiconductor has recently introduced the 11C00 family of sub-nanosecond logic for instrumentation applications. The entire family is voltage compensated to improve noise margins and eliminate the ±2% power-supply regulation requirements of uncompensated ECL ICs, thus making system design much more simple. Isoplanar II processing is used to achieve maximum speeds and keep die size to a minimum. At this writing the logic family consists of only about six devices, but two of these should be of immediate interest to the amateur.

GHz prescaler

The 11C05 is an asychronous divide-by-four counter which operates from a single power supply and features toggle rates in excess of 1000 MHz over the entire 0°C to +75°C temperature range. The input may be ac or dc coupled so that either an input amplifier or a simple biasing network may be used. The single rail power requirement allows the use of a +5 volt supply in predominately TTL systems. The 11C05 will toggle with a sinusoidal input to a minimum of about 25 MHz, or, with a square-wave input having fast edge rates, to dc. A circuit showing the 11C05 in a divide-by-forty uhf prescaler appears in fig. 1. The many owners of existing 95H90 vhf prescalers will be pleased to note that the 11C05 can be directly interfaced to their present circuits with a minimum of modifications.

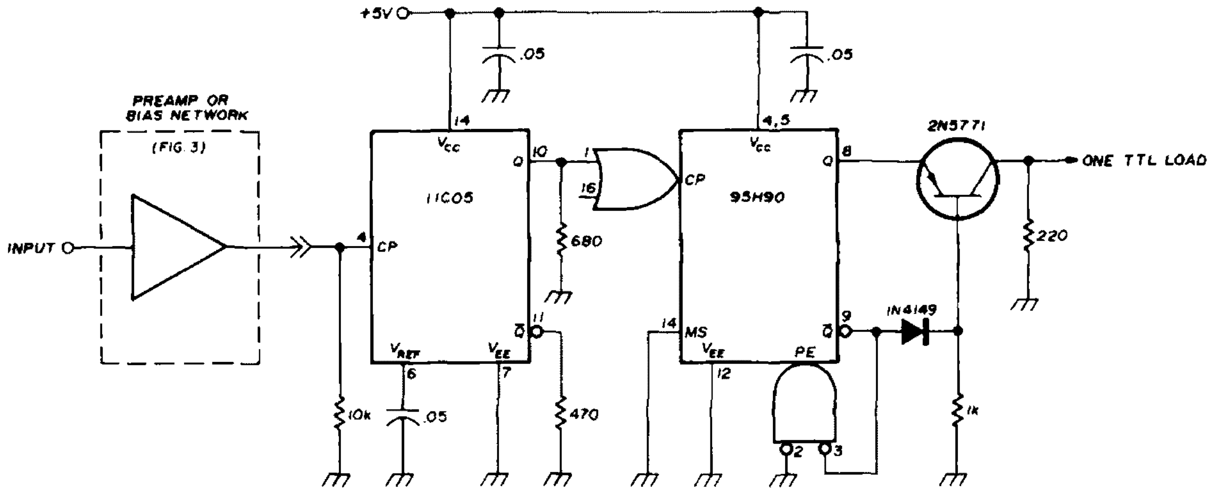

fig. 1. Divide-by-40 prescaler for use to above 1000 MHz uses 11C05 divide-by-4 counter and 95H90 decade divider (unused CP input is tied to ground). Ac-coupled 11C05 input requirements vs frequency are plotted as shown in fig. 2.

As shown, with an input amplifier (an Amperex ATF417 was used in one design), the input sensitivity is better than 50 millivolts rms to above 1000 MHz. The 10k resistor from pin 4 to ground is included to eliminate noise triggering in the middle frequency ranges. A glance at fig. 2, a plot of input sensitivity vs frequency, shows that this is necessary. Fig. 2 also shows that no input amplifier is necessary between about 150 and 800 MHz if a minimum-cost design is desired. The function of the remaining termination and bypassing components associated with the 11C05 is obvious.

The balance of the prescaler consists of the 95H90 in a standard configuration with the 2N5771 used as an ECL-to-TTL level translator capable of driving one unit load. Although the diode used from the 2N5771 base to the $\bar{Q}$ output of the 95H90 was a low-capacitance 1N4149 type, the more common 1N4148 should perform satisfactorily.

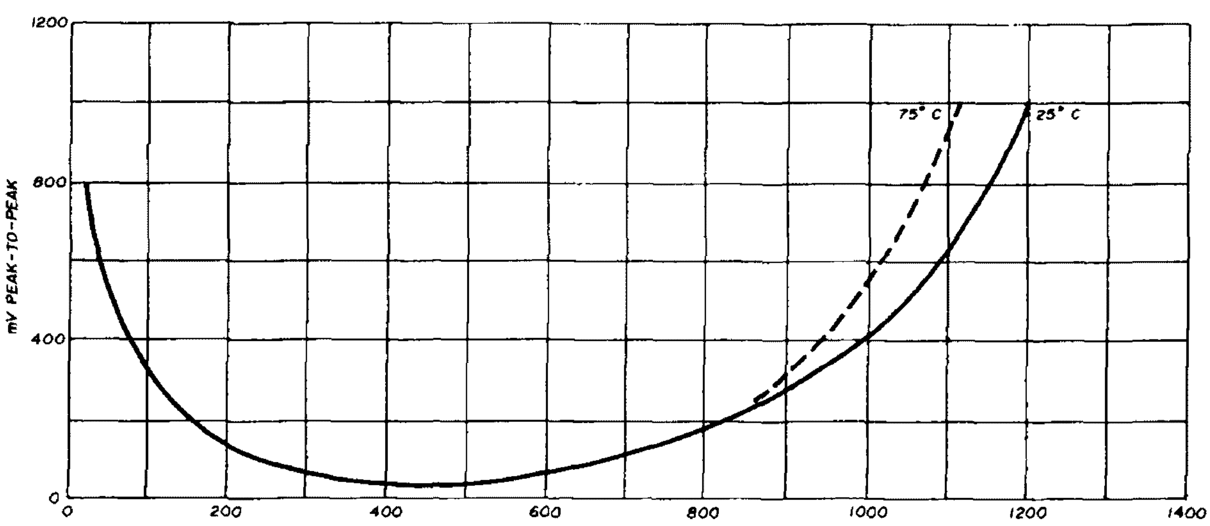

fig. 2. Ac-coupled input requirements for the Fairchild 11C05 1000-MHz divide-by-4 counter.

The less popular 95H91 divide-by-five counter may be directly substituted for the 95H90 with no change in wiring. This results in a divide-by-twenty prescaler suitable for use with counters capable of 50-MHz input frequencies and offers some additional advantages, e.g., mental multiplication of the counter reading by two is easier than multiplication by four. Also, a simple timebase divider can be built to double the gate time and yield the correct display.

uhf prescaler

The 11C06 is a 700-MHz type-D flip-flop. When used as shown in fig. 3, a divide-by-twenty uhf prescaler with toggle rates in excess of 550 MHz from 0°C to +75°C is the result. Again, an amplifier or an input biasing scheme may be used. An unamplified input was chosen for this design to illustrate its simplicity and also its adaptability to the 11C05. This circuit may also be used with existing 95H90 designs, or a 95H91 could be substituted to build a uhf decade prescaler that eliminates the need for mental gymnastics or time-base modifications. The only concern is that the counter must accommodate the desired maximum frequency divided by ten.

operation

As noted above, when either of these prescalers is used with most counters, it will be necessary to multiply the reading by the scale factor or to build a simple binary divider and insert it into the circuit between the crystal and the existing divider chain. This divider should then be switched into operation anytime the prescaler is in use. An unused set of contacts on this switch could be used to reposition the counter's decimal point one place to the right to reconcile the decade of division. The only drawback to the time base extension is the doubling of the gate time which results in a two-second gate and a four-second wait for display update. However, to most people, the correct display is probably worth the wait.

No 1296-MHz signal source was available for testing, so it is not known whether the 11C05 prescaler can be used at this frequency. However, it appears that some devices would operate successfully, particularly at lower temperatures.

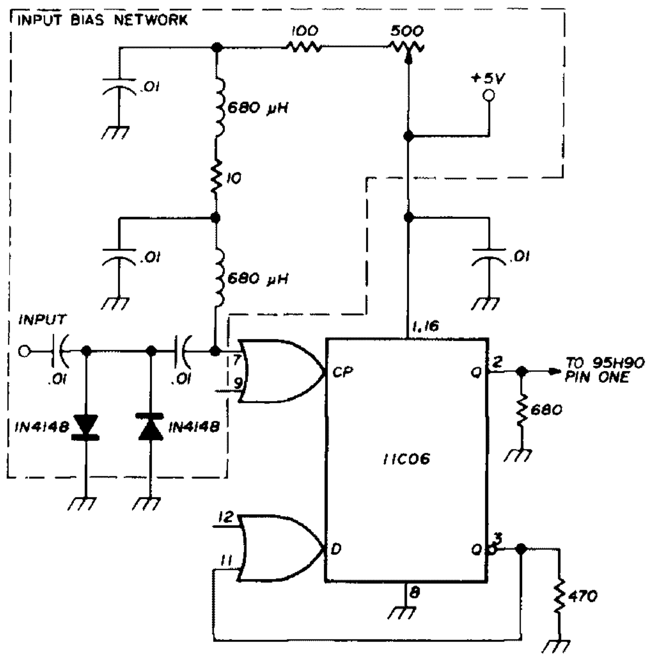

fig. 3. The Fairchild 11C06 700-MHz type-D flip-flop may be used with a 95H90 decide counter to build a divide-by-20 prescaler, (unused CP and D inputs are ties to ground).

conclusion

In summary, the Fairchild 11C00 series of sub-nanosecond logic has extended the range of digital techniques beyond 1000 MHz. As additional elements such as phase-locked loops, wideband amplifiers, phase and frequency comparators, etcetera, are added, the applications base will broaden to include communications, frequency synthesis and data handling. In small quantities the 11C05 is $87.90 and the 11C06 is $21.97; both are available through franchised Fairchild distributors.

ham radio

fm channel scanner

for the Heathkit HW202

Adapting K2ZLG's popular vhf scanner circuit to the Heath HW202

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The scanner described by K2ZLG in the February, 1973, issue of *ham radio* can be built into the popular Heathkit HW202 vhf transceiver at a cost of about ten dollars, making a fine addition to the unit. It is extremely useful when mobiling in areas of light two-meter activity and saves wear and tear on the channel-selector pushbuttons (as well as the operator's fingers).

The circuit in **fig. 1** is essentially the same as that described by K2ZLG. However, all parts not required for operation with the HW202 have been eliminated. One unexpected but fortunate parts saving resulted from the discovery that power can be supplied to the scanner by connecting it in series with the lamp that illuminates the panel meter. This results in less battery drain than when the circuit is powered by a separate voltage-dropping resistor and zener diode. Somewhat less than the optimum five volts is supplied to the ICs, but due to the relatively slow toggling speed, no problems have been encountered.

The lamp behind the meter, however, blinks a bit with the scanner's fluctuating load; if this bothers you, a zener-regulated supply can be installed. To operate the receiver in the scan mode, it is only necessary to turn on the scanner power and depress one of the channel selector buttons part way so that all three are unlatched. If you wish to manually select one of the channels, power to the scanner must be removed so that two crystals cannot be switched in at the same time.

construction

The scanner is designed to fit into the space normally occupied by the optional Heath tone-encoder kit. If you have already installed the tone encoder you might be able to mount the scanner externally. However, it would be better to consider outboarding the tone encoder due to the number of leads involved with the scanner circuit.

toggle switch with a shank long enough to go through one of the holes in the sub-panel and also reach through the plastic strip.

Another amateur who modified his unit obtained a small slide switch from a hobby shop which worked as well and mounted nicely on the plastic strip. Either way, a little care is required to get everything to fit together properly

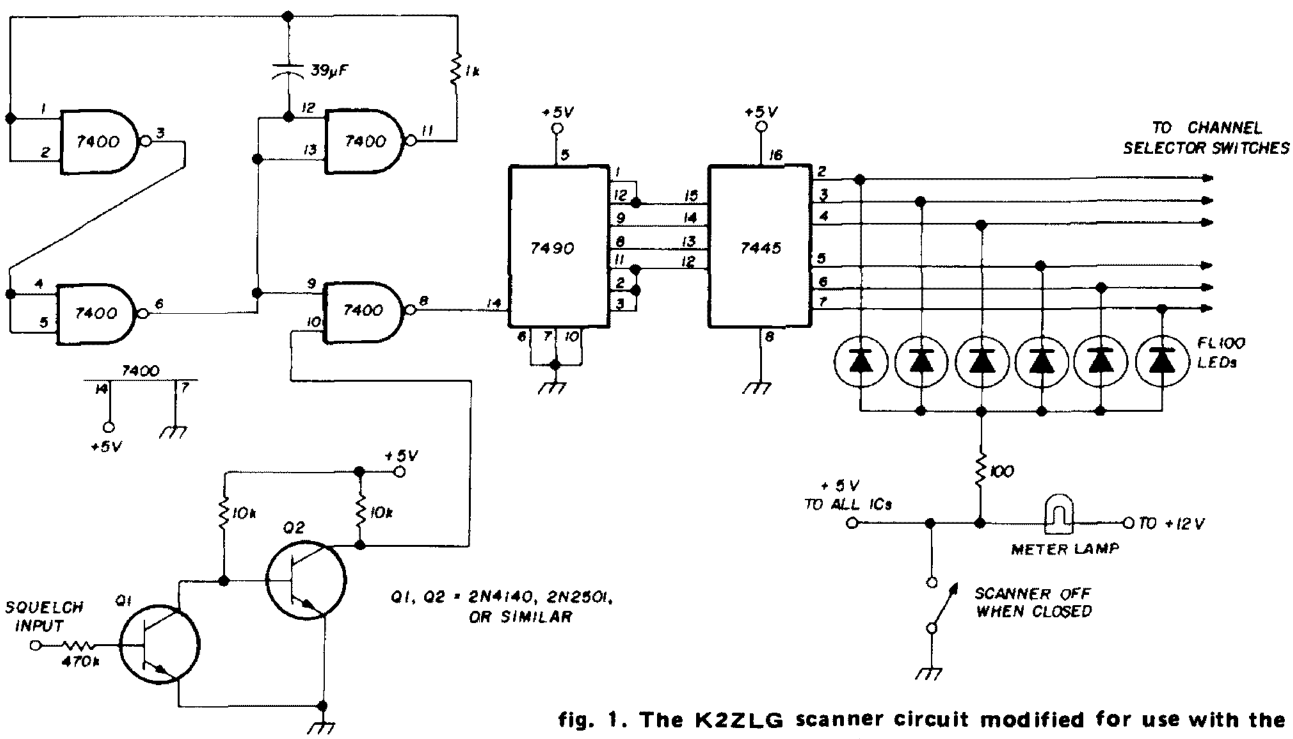

fig. 1. The K2ZLG scanner circuit modified for use with the Heath HW202 fm transceiver.

The parts were assembled on a printed-circuit board which was tested for correct operation before permanently mounting it inside the transceiver. There is nothing critical about it, and hard-wired or perf-board construction will work as well as printed circuitry.

The LEDs and the scanner on/off switch were mounted on the plastic filler plate which was removed from the front panel. Holes were pierced in the plastic strip and the LED leads were run through the holes and extended through a hole in the sub-panel with color-coded wires. The color code was matched to similar wires connected to the channel switches. The on/off switch is a small

without shorting any of the LED leads during assembly. Be sure to observe polarity when connecting the LEDs.

It is necessary to partially remove the front panel for easy access to the receive channel pushbuttons. After you get to them, connect a color-coded wire to each of the two unused lugs at the bottom rear of each of the three channel buttons. Then bring all six wires out to the side of the chassis and reassemble the switches and the front panel. It is helpful during the debugging process if you make these wires fairly long. They can be easily shortened later.

The scanner input lead is connected to the base circuit of Q107, the squelch

emitter follower. The upper end of R174, an 82k resistor, is a convenient point to make this connection without removing the receiver circuit boards.

As mentioned before, power is obtained from the lamp circuit. This is easily done by removing the ground connection from the base of the lamp and connecting the positive input of the scanner board to this point.

testing

When all the connections have been made and power applied to the transceiver, the LEDs should light in sequence from left to right. Manually opening the squelch, or an incoming signal on any channel, should stop the searching action. You may have to adjust the value of the 470k resistor in the squelch input circuit slightly, depending on the gain of the transistors used on the scanner board. The rate at which the scanner operates can be adjusted by changing the value of the timing capacitor in the 7400 clock circuit. The 39-μF value results in a scan rate of about six channels per second.

circuit noise

After getting the scanner in operation, you may notice a clicking noise that can be quite annoying, especially in a quiet room. These clicks are simply key clicks from the switching oscillator stage in the receiver and can be reduced by connecting a 2700-ohm resistor from the base of transistor Q116 to ground (the oscillator stage). The resistor will reduce the transistor's forward bias during the time the scanner is moving from one channel to another. In addition, the emitters of transistors Q109 and Q110 are returned to ground through R183 which is grounded to a portion of the PC foil shared by the oscillator stage. A 680-pF capacitor from the emitters of Q109 and Q110 to the nearest ground on the circuit board should completely eliminate the clicks.

ham radio

a second look (from page 4)

while others are opposed because a large number of Communicators on our vhf bands would create undesirable crowding and interference. Remember, however, that our vhf allocations are very susceptible to raiding by other radio services — a healthy and growing amateur population is probably the most effective weapon we have.

If the proposed regulations are adopted, the written examination for the Advanced Class will apparently be essentially that which is now required for the Amateur Extra Class. If you've been thinking about going for your Advanced ticket anyway, perhaps now is the time to consider it seriously because the present examination is probably considerably easier than the new one will be. And, if the new regulations are adopted, the only privileges the Advanced Class will be denied are the small high-frequency CW segments which will continue to be reserved for the Extra Class.

Under the new scheme, incidentally, the Extra Class exam will consist only of a 20 wpm telegraphy test. This class will carry *both* high-frequency and vhf privileges, and will be issued for life (only the station license need be renewed every five years). The new regulations will also give Technicians full six- and two-meter privileges (50.0-50.1 and 144-145 MHz), so nearly everybody gains.

Docket 20282 is probably the most far-reaching proposal affecting amateur radio which has ever been issued by the FCC, and it deserves the attention of all of us. Since copies of the complete Docket have been sent to all the subscribers of *ham radio*, and the ARRL has sent copies to all their affiliated radio clubs, the material is widely available. Read it, then stand up to be counted. Comments are due at the FCC by June 16, 1975.

Jim Fisk, W1DTY
Editor-in-Chief

transistor

breakdown voltages

A complete discussion of transistor breakdown voltages, what the ratings mean, and how they affect the application of the device

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The pages of all the electronics magazines are presently filled with lucrative offers of all types of semiconductor devices, and many of the prices are really quite reasonable! Most amateurs operate on a limited budget anyway, so these offers are quite tempting. The ads often state "... replaces types ..." but all too often this is not the case. Why?

One of the major stumbling blocks to a thorough understanding of the applications of transistors is the lack of a working knowledge of transistor breakdown voltages. Breakdown in a device can often lead to its sudden and mysterious failure without leaving any trace as to what happened. Another similar device is then inserted, and the same thing may or may not happen. It's no wonder that the amateur who does his own design or substitutes "grab bag" goodies into a proven circuit is often mystified when the circuit fails to operate properly.

diodes

In its simplest form, breakdown in semiconductor devices can be observed in an ordinary diode. The diode is a junction of positive (p) and negative (n) type semiconductor materials, and theoretically will conduct current in only one direction. An illustration of this forward bias condition is shown in **fig. 1**. The relationship between the current i and the voltage v will be defined later.

If the diode is inverted in the circuit no current (except for reverse leakage current) will flow. This will remain true until the reverse voltage reaches a point where the current begins to increase almost without limit. The value of voltage at which this phenomenon occurs is known as the breakdown voltage. **Fig. 2** shows the "reverse bias" configuration which will lead to eventual breakdown.

Depicted in **fig. 3** is a graphical illustration of the relationship between diode voltage and current. In the forward bias region, current tends to in-

crease rather smoothly as voltage is increased. In the reverse bias condition, however, current flow seems to remain at some small and almost constant value until breakdown occurs; reverse current then increases very rapidly. The reverse current is referred to on data sheets as "reverse saturation current"; symbolically, it is called I_O. Both I_O and BV vary considerably among individual devices, and ranges of values are sometimes presented on data sheets along with other pertinent characteristics.

It should be noted that the well known and widely used zener diode is actually a diode used in the reverse bias region. Its breakdown or *zener* voltage is closely controlled and specified.

In diodes, the reverse breakdown voltage is sometimes called "peak inverse voltage," or PIV. In the selection of rectifier diodes, proper attention to PIV is always given for the potential of breakdown is quite clear.

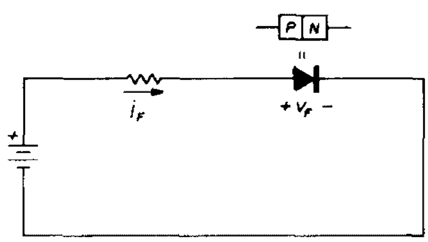


fig. 1. Forward-biased semiconductor diode.

transistors

Now that some insight into a diode's behavior has been given, transistors can be tackled! Consider the circuit of **fig. 4**. In this particular case the collector-base junction can be treated simply as the ordinary diode just discussed. As before, a reverse leakage current (now called I_{CO}) flows through the collector-base junction and some value of reverse bias will eventually be reached where I_{CO} begins a rapid increase. This, again, is the value of the breakdown voltage, and is designated as BV_{cbo}. Stated another way, it is the breakdown voltage between collector and base with the emitter lead unconnected ($I_e = 0$).

Even if some emitter current is allowed to flow, the breakdown point will still approach BV_{cbo}, but the current increase is not as dramatic as when I_e is zero. **Fig. 5** will clarify this. A similar condition exists whenever the base terminal is left unconnected. This time, however, the voltage of importance is called BV_{ceo} (see **fig. 6**). Note that BV_{ceo} is less than BV_{cbo}. The actual difference is dependent upon the electrical parameters of the individual devices.

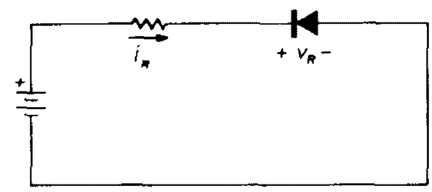


fig. 2. Reverse-biased semiconductor diode.

Strictly speaking, the open-circuit base is not often encountered, so some alterations should be considered. Connecting the base to the emitter through a resistor modifies the I_c/V_{ce} relationship as evidenced by **fig. 7**. With the resistor in the circuit, breakdown can be increased above BV_{ceo} to BV_{cer}. If the resistor is actually decreased to zero ohm, another value, BV_{ces} (the "s" is for short circuit) is obtained.

As if this weren't enough to remember, there is still one more breakdown voltage — BV_{cex}. This is generated by applying negative bias to the base. For

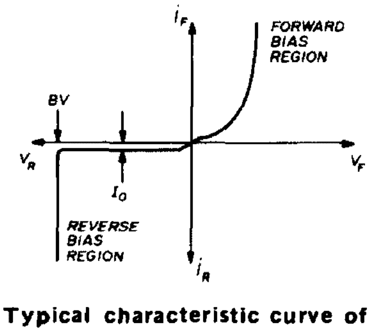


fig. 3. Typical characteristic curve of a semiconductor diode, showing operation in the forward- and reverse-biased regions.

the purposes of this discussion it is not necessary to know the empirical relationships between BV_{ceo}, BV_{cer}, BV_{ces} and BV_{cex}. However, knowing their approximate relationships to each other and under what conditions they become important should be one of your goals. Reference to **fig. 7** should provide some insight into approximate magnitudes of the various breakdown voltages.

From this graph it can be seen that breakdown can range in values from a low of BV_{ceo} to a high of BV_{cbo}. By no means do these curves represent all transistor types; they are, in fact, somewhat idealized. In spite of this, they will give you a feel for the concept of breakdown.

using the data

At this point some attention should be given to the application of the

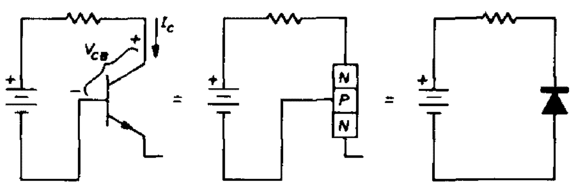


fig. 4. Diode equivalent of a transistor's collector-base junction.

various ratings just discussed. Each rating has its own special significance, and it would hardly prove practical to discuss each in detail. What will be useful, though, is the formulation of some general guidelines to help you in the application of the transistors themselves.

While it is unusual to find a wide variety of breakdown voltages specified on a data sheet, BV_{cbo} is commonly specified. In order to wisely apply a device, however, BV_{ceo} should also be known since it establishes the minimum breakdown voltage. Another item usually stated on data sheets is the static forward current transfer ratio, or dc beta; its designation is h_{FE}. The actual value of BV_{ceo} is dependent upon these two items just mentioned. It can, in

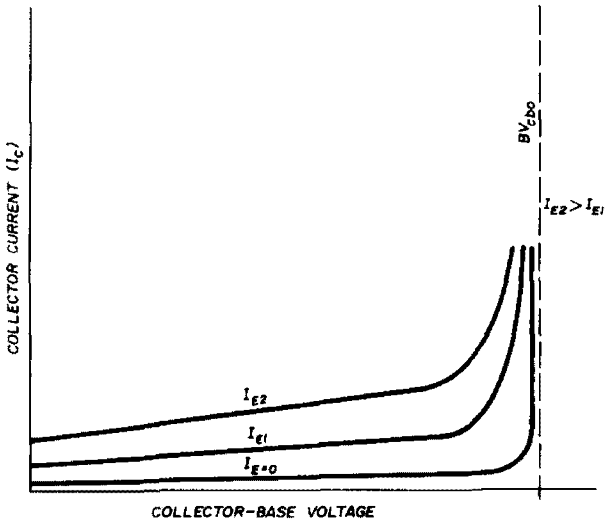


fig. 5. Transistor collector-base breakdown characteristics with the emitter open.

fact, be approximated from the equation

$$BV_{ceo} = BV_{cbo}/(1+h_{FE})^{1/n}$$

where n is an experimentally determined factor. For purposes of approximation, you can assume n to have a value of 2.5 for silicon transistors and a value of 6.0 for germanium transistors. **Fig. 8** is included to aid in the estimation of values for BV_{ceo}.

From the equation you can see that for a current gain of zero, BV_{ceo} will equal BV_{cbo}. With small-signal devices,

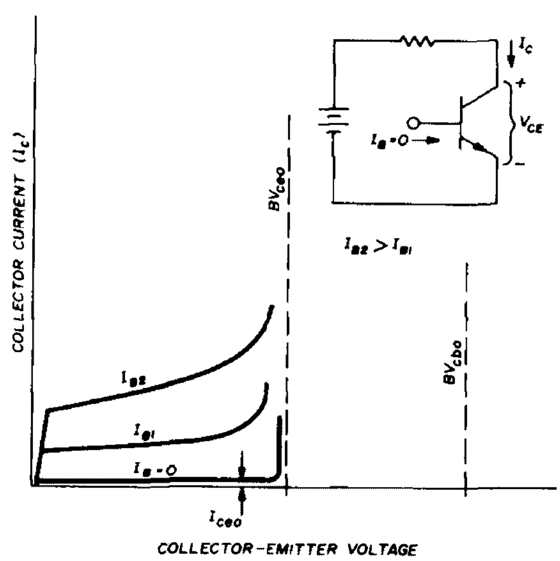


fig. 6. Transistor collector-emitter breakdown characteristics with the base open.

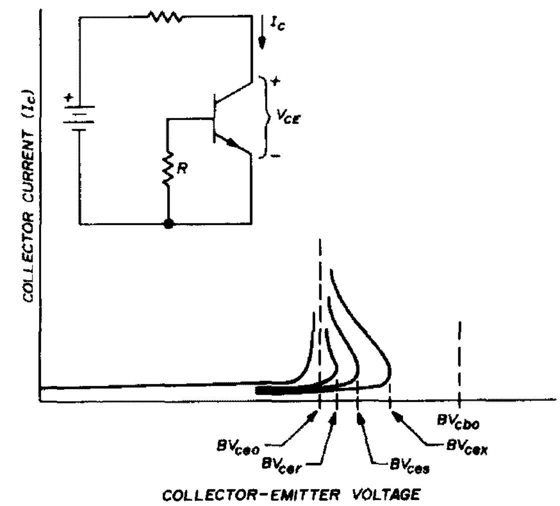


fig. 7. Transistor collector-emitter breakdown characteristics for various base biasing conditions.

current gain is usually rather large, so BV_{ceo} will, in most cases, be the smaller number by a rather wide margin. This has the practical significance of a built-in safety factor – use of the BV_{ceo} rating for selection of a device will almost always insure that breakdown ratings will not be exceeded because the base lead will probably not be anywhere near an open circuit. For additional reliability, though, an additional safety factor may be applied. In many military designs this factor will vary from 0.5 to 0.75. For amateur work, 0.75 will suffice.

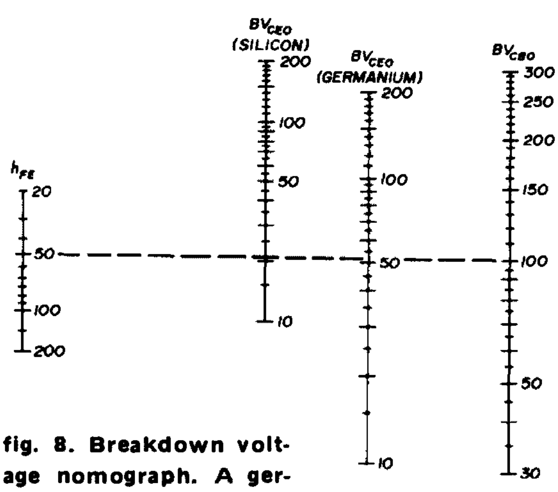


fig. 8. Breakdown voltage nomograph. A germanium transistor, for example, with a forward current gain (hfe) of 50 and BV_{cbo} of 100 volts has a BV_{ceo} of approximately 52 volts (if the transistor were silicon, the BV_{ceo} would be slightly more than 20 volts).

The term "breakdown" carries with it the connotation of complete destruction, but this is not always the case. Device destruction often results because large voltages and significant current are present in a device simultaneously, and its power rating is exceeded. In many instances, however, breakdown is used to advantage. Current flow is limited to the extent that the product of current and breakdown voltage falls well within the dissipation rating for the device.

As previously mentioned, the zener diode is one application of controlled breakdown. The emitter-base junction of transistors (the breakdown voltage

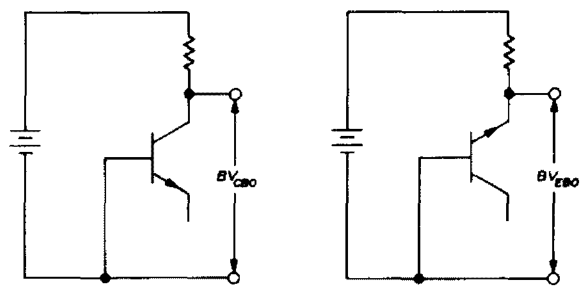


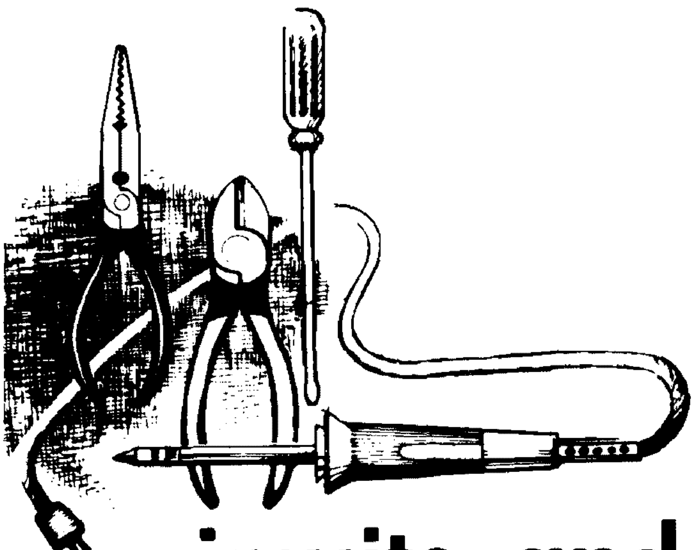
fig. 9. Using transistor junctions as zener diodes.

ranges from 0.5 to tens of volts, depending upon the device type) is often used as a zener diode. Similarly, the collector-base junction is sometimes connected for use as a high-voltage zener (voltages up to several hundred volts are possible). **Fig. 9** illustrates these two common applications.

conclusion

From this discussion it should be clear that the term "breakdown voltage" does not refer to a single, well-defined quantity. Likewise, the process of breakdown in semiconductor devices is not an easy one to comprehend. While the presentation here is not intended to be exhaustive, the information should allow a reasonable approach to device selection and substitution.

ham radio



circuits and techniques

ed noll, W3FQJ

mosfet circuits

The metal-oxide semiconductor field-effect transistor (mosfet) is now used extensively in transmitters, receivers and test equipment. Some common and some unique circuits have been built around this device and are being used in both amateur and commercial communications equipment. They perform well as ampli-

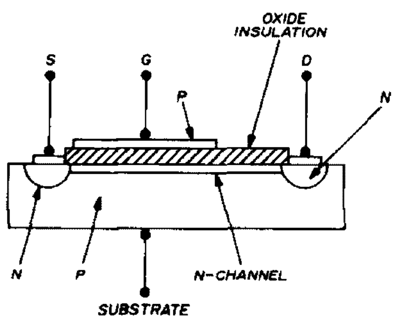


fig. 1. Construction of the depletion-type mosfet.

fiers up into the high uhf spectrum, and the dual-gate types are particularly popular in mixer, oscillator and converter circuits. In addition, they serve ideally as balanced modulators and demodulators.

Basically, the mosfet differs from the junction fet in that the gate itself does

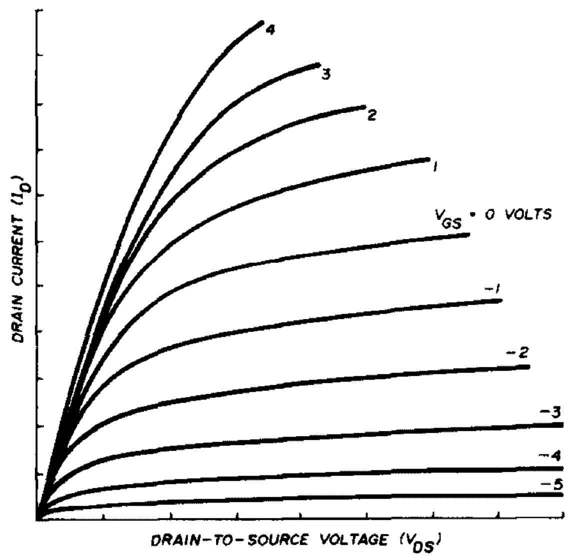


fig. 2. Operating characteristics of the n-channel, depletion-mode mosfet.

not actually touch the channel — there is an intervening metallic-oxide layer between gate and channel as shown in **fig. 1.** However, the gate electrode acts as a control element just as it does in a junction fet. The oxide insulation between the gate and the channel keeps the leakage current very low and the input impedance very high. Nonetheless, the charge placed on the gate determines the charge motion along the channel between the source and drain. As in the case of the junction fet, the gate charge determines the extent of the depletion region in the n-channel.

As the gate is made negative, relative to the source, the number of electrons in the channel is depleted. This activity is similar to that of a junction fet. During normal operation the gate of the junction

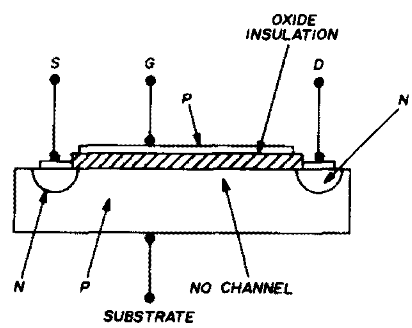


fig. 3. Construction of the enhancement-mode mosfet.

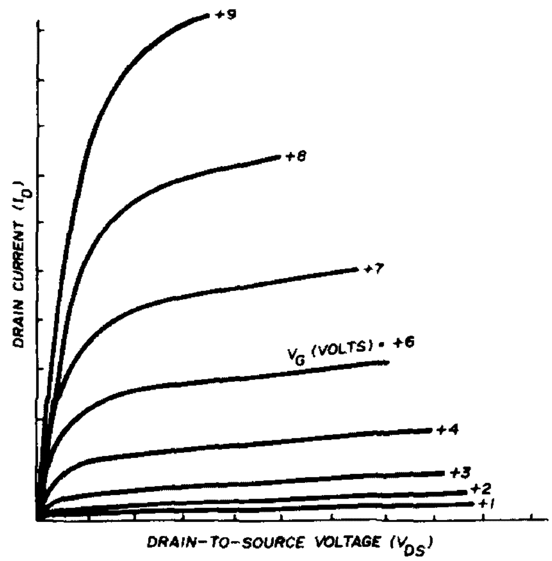


fig. 4. Operating characteristics of the n-channel, enhancement-mode mosfet.

fet may not be permitted to swing significantly past zero voltage because the junction will then be forward biased and input impedance falls. The mosfet has no such limitation; the gate can be permitted to swing past zero, increasing channel conductivity without any increase in the gate current or drop in input impedance.

The drain current vs drain voltage characteristic of a depletion-type mosfet is plotted in **fig. 2**. Note that the gate signal swings to either side of the zero-volt gate-bias curve. Drain current rises on the positive side; it falls on the negative side.

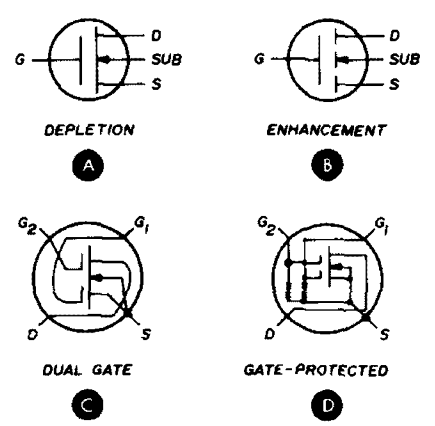


fig. 5. Symbols used for various mosfets. N-channel, depletion type is shown in (A); n-channel, enhancement-mode type is shown in (B). (C) shows dual-gate, depletion-mode device while (D) shows a diode-protected mosfet.

A second basic type of mosfet is the enhancement-mode device shown in **fig. 3**. In this device there is no channel present in the substrate that exists between the n-type source and n-type drain. With no bias applied to the gate there is no channel current. Likewise, a negative gate charge results in no channel current.

If a positive bias is placed on the gate of an enhancement-type mosfet it will attract electrons from the p-substrate into the region beneath the gate, forming an n-type channel that links the n-type source with the n-type drain. Consequently, there is a charge motion along

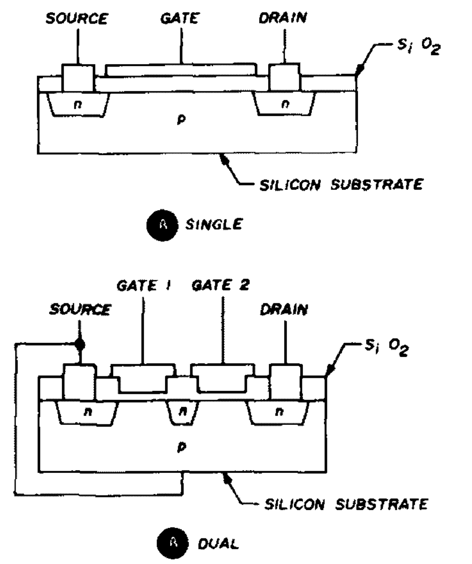


fig. 6. Construction of single-and dual-gate mosfets.

the channel which increases with the positive charge applied on the gate. If the positive bias voltage is made to vary with signal there will be a like change in the channel current. The I_D vs D_{DS} curves of **fig. 4** show how the drain current increases with an increase in the positive gate bias.

The basic symbols for the different mosfet types are shown in **fig. 5**. These symbols signify an n-type channel. However, they can also use a p-type channel – the symbolization is identical with the exception that the substrate arrow must be pointed in the opposite direction. The symbol in **fig. 5A** refers to a depletion-

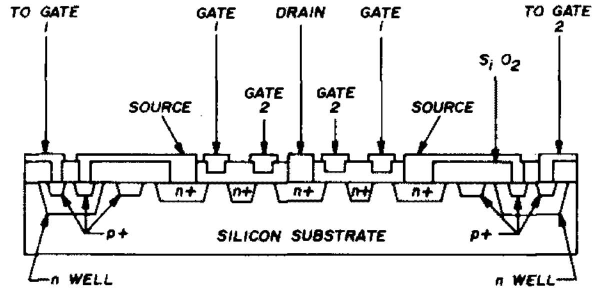


fig. 7. Dual-gate mosfet with diode gate protection.

type n-channel mosfet; **fig. 5B** refers to the enhancement-mode type. **Fig. 5C** and **5D** refer to dual-gate, depletion-type mosfets. **Fig. 5D** refers to mosfets that include internal protective diodes.

The high input impedance is both an advantage and a hazard of the mosfet device. Since the input impedance is very high, a tiny current or static charge can build up a very high level on the gate, possibly destroying the device. Thus mosfets must be handled carefully and circuit arrangement must be such that excessive charges are excluded from the gate(s). The internal protective diodes avoid these hazards by establishing conducting paths when the gate charge becomes excessive.

dual-gate mosfet

The dual-gate mosfets bring added versatility to mosfet circuit design. The two gates and balanced configuration is attractive for all types of balanced amplifiers, modulators, mixers and demodulators. In straight amplifier and mixer applications the second gate provides a means of applying agc voltage or local-oscillator injection. Most dual-gate mosfets are n-channel depletion types, **fig. 6**. The two devices shown are similar except for the addition of a second gate. As a result there are two electrodes that control the conductivity of the channel. In fact, in mixer circuits the signal applied to gate 2 is used to modulate the transfer characteristics of the input gate. As a result there is a form of mixing that has improved linearity over conventional square-law devices. Good conversion gain is obtainable with minimum injection level.

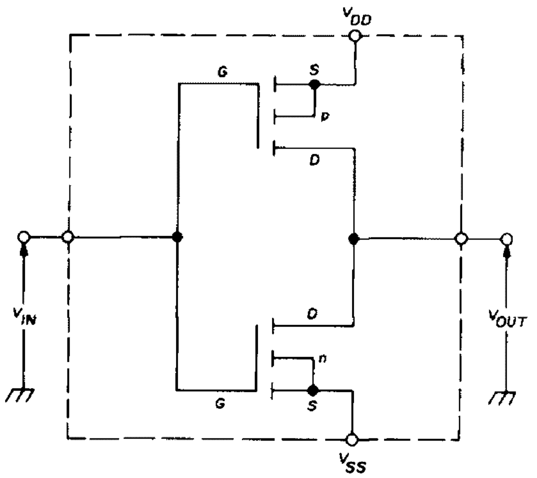


fig. 8. A complimentary-symmetry/metal-oxide (cmos) circuit.

The addition of internal protective diodes protects the gates against damage from normal handling and use. These diodes drain off high-voltage charges and protect the input circuit from excessive voltages and signals. For example, any voltage transients in excess of ±10 volts are bypassed by the back-to-back diodes connected between each gate and the

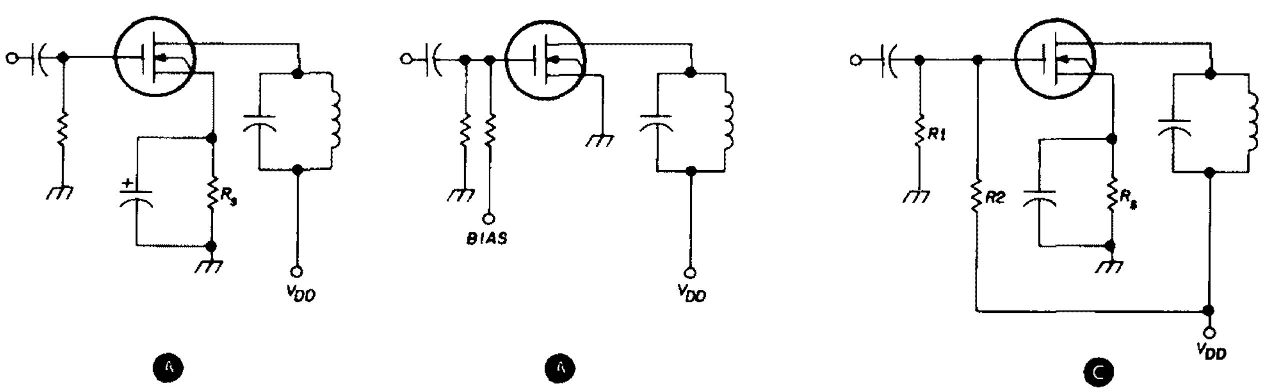


fig. 9. Biasing circuits for single-gate mosfets.

substrate. As shown in the basic structure of the gate-protected mosfet chip, **fig. 7**, there are two n-type input wells. The p-region of the gate forms a diode junction with the well, the well junction with the source serving as the second diode. It is interesting that such a protected-gate mosfet is less subject to static discharge damage than even a bipolar transistor.

Mosfets lend themselves to inclusion in integrated circuits. One of these is the popular cos/mos or cmos IC using both enhancement and depletion types. The term cos/mos (cmos) refers to complimentary-symmetry/metal-oxide semiconductor devices. Two enhancement-type mosfets are shown in the cmos IC in **fig. 8**. Note from the direction of the substrate arrows that the upper device has a p-channel and the bottom device an n-channel. Such configurations are especially adaptable to logic circuits incorporating the attractive features of high input impedance (low input capacitance) and very low power demand.

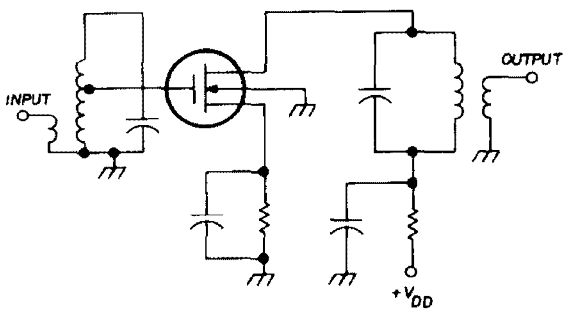


fig. 11. Simple rf amplifier circuit using a single-gate mosfet.

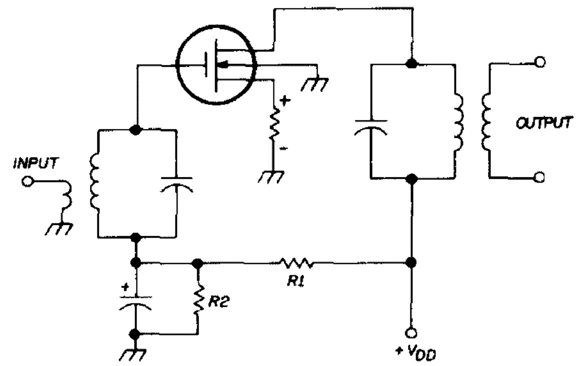


fig. 12. Mosfet rf amplifier circuit using resistor-divider biasing.

Biasing methods for the single-gate mosfet are given in **fig. 9**. The insulated gate provides a high input impedance and, like the pentode vacuum tube, there is no significant input current. The circuit in **fig. 9A** is biased with the source resistor R_s, the direction of source current being such that a positive bias is developed in the source circuit. This in effect biases the gate negative for the n-channel, depletion-type mosfet. Operation is similar to the cathode biasing of a vacuum tube. If degenerative feedback is wanted, the filter capacitor is eliminated from the source circuit.

External biasing, using a two-resistor divider, is shown in **fig. 9B**. In this arrangement the source and substrate are both grounded. The most popular form of mosfet biasing uses a combination of both types as shown in **fig. 9C**.

In circuit application there can be a considerable variation in drain current for individual devices. Thus, with a certain fixed bias, the actual drain current could fall between somewhat wide limits for a

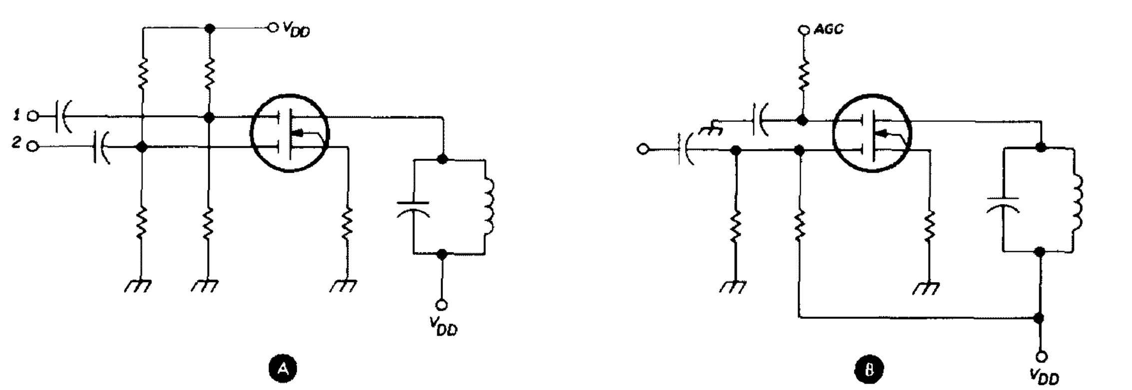


fig. 10. Biasing methods for dual-gate mosfets.

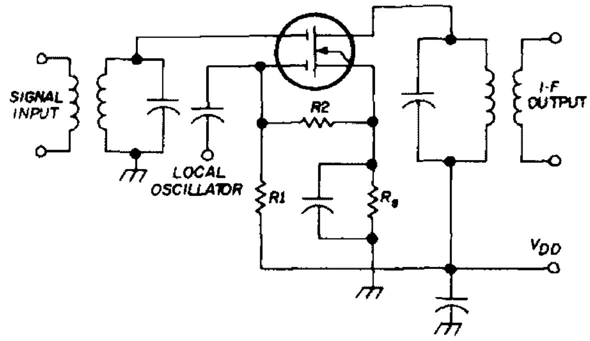


fig. 13. Mixer circuit using a dual-gate mosfet.

given transistor type. The combination of external divider bias and source resistor bias confines this limit to a much narrower range.

Dual-gate biasing schemes are shown in **fig. 10**. In **fig. 10A** a combination of two-resistor divider and source resistor bias are used. Note that there are two inputs, suggesting that the circuit could be used as a receiver mixer or for signal combining or switching. The circuit in **fig. 10B** shows a basic rf amplifier stage using

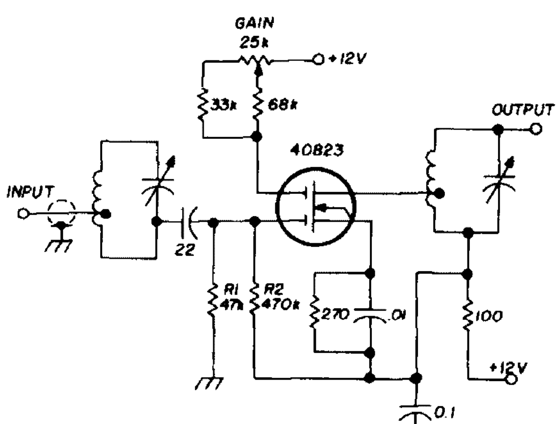


fig. 14. High-frequency rf amplifier using a dual-gate mosfet.

similar biasing. The exception here is that one gate is used for agc voltage or some other form of controllable bias, as from a receiver rf gain control.

basic circuits

A typical weak-signal rf amplifier using a single-gate mosfet is shown in **fig. 11**. In this circuit a small amount of source bias is used; the substrate itself is grounded.

When greater dynamic operating range is required and/or critical operating conditions must be maintained, a combination of source and external bias can be used (**fig. 12**). The way in which the bias dividers are connected is such that there is minimum loading of the input resonant circuit. Instead, the bias voltage is developed across the capacitor connected between the bottom end of the resonant circuit and common. Therefore, it is not necessary to shunt the divider resistors between the high-impedance input gate

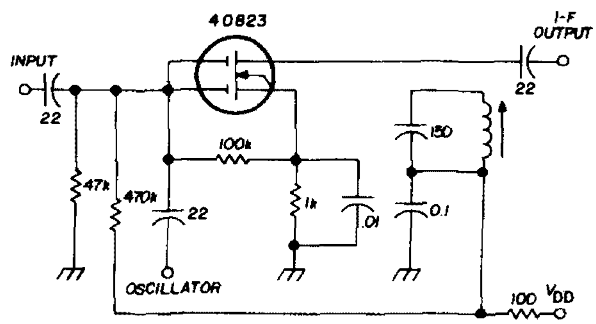


fig. 15. Mosfet dual-gate mixer.

and common as is the case for the bias arrangement shown in **fig. 9C**.

In the combination-bias arrangement it should be noted that negative gate bias is developed across the source resistor while the two-resistor divider results in a positive bias voltage. The actual gate bias is the algebraic sum of the two.

A very popular mixer circuit for modern receivers is shown in **fig. 13**. The signal is applied to the top gate; its bias is determined largely by the source resistor R_S. Local-oscillator injection is made at the lower gate. Optimum mixing bias is established by R1/R2 combination.

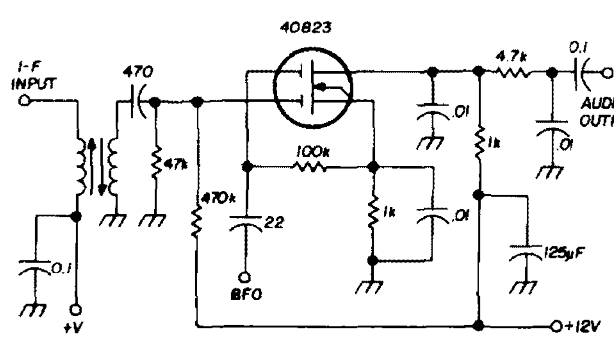


fig. 16. Product detector. Excellent isolation is provided by the dual-gate device.

receiver circuits

The circuits that follow are very practical ones gathered from proven amateur and commercial radiocommunication equipment. Parts values are given whenever they are known. The first three circuits are receiver types used in the Ten-Tec Argonaut transceiver. They demonstrate how a dual-gate depletion-type mosfet can serve in a number of basic receiver circuits.

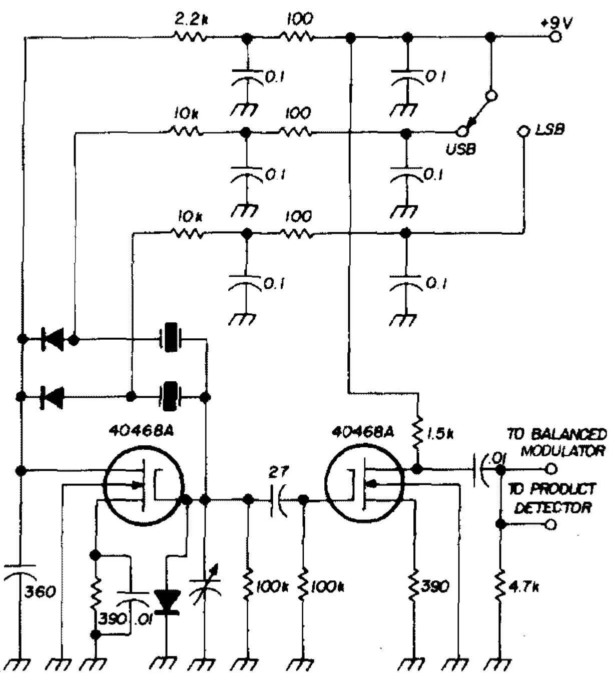


fig. 17. Sideband carrier oscillator using diodes for upper/lower sideband switching.

A high-frequency amplifier is shown in fig. **14**. The input circuit has high Q and the low impedance of the antenna is matched through the tap at the low end of the coil. The input gate tap is further up the coil. A resonant output system is included in the drain circuit. In application this can be a high impedance output or, if a low impedance output is desired, it can be obtained by tapping at the low end of the output coil.

Input gate biasing is handled with divider resistors **R1** and **R2**. The second gate is used to control the gain of the rf amplifier. Note the convenient manner in which this can be accomplished using two

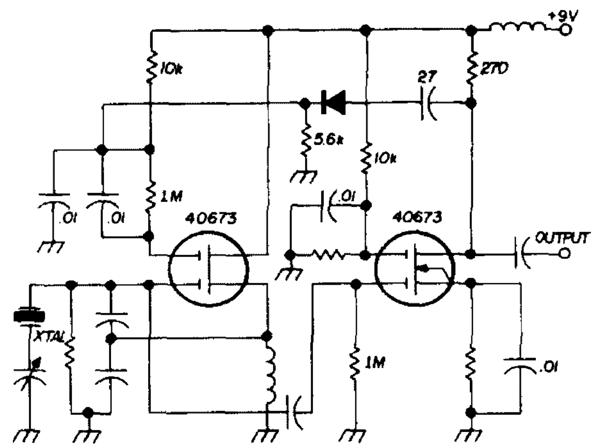


fig. 18. Wide range, untuned, high-frequency crystal oscillator.

fixed resistors and the gain control potentiometer.

A signal mixer circuit is shown in fig. **15**. Separate gates provide input for signal and local-oscillator injection. Component values are similar to those of the rf amplifier except for the simple gate-input circuit for the local oscillator signal. The output resonant circuit is tuned to the intermediate frequency.

Fig. 16 shows the same mosfet being used as a product detector for demodulation of CW or ssb signals. Note the many similarities of the three receiver circuits. Such a device and associated components could be easily set up on a vector board if you wanted to try a bit of mosfet experimentation.

The i-f signal arrives by way of the input resonant circuit. The demodulating oscillator supplies signal to the second gate. High-frequency components are filtered out in the drain output circuit. A

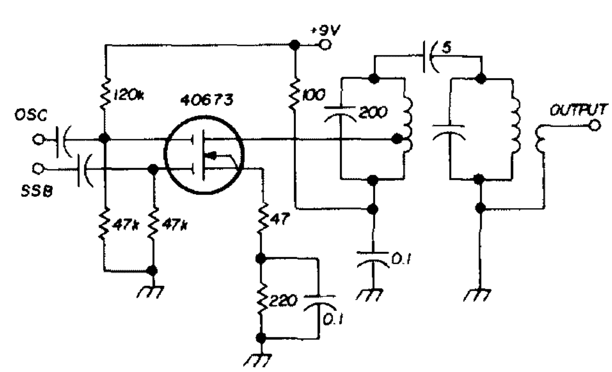


fig. 19. Sideband mixer circuit using a dual-gate mosfet.

resistor-capacitor lowpass filter passes the voice frequencies to a succeeding audio amplifier.

The next four circuits were gleaned from the sideband transceiver manufactured by Sideband Associates (SBA) for operation in the 2- to 23-MHz range and used primarily for radio-marine communications. Two mosfet devices are used in the carrier oscillator, **fig. 17**, along with a diode switching arrangement that can be used to select either the upper or lower sideband. The 40468A is a single-gate device. The circuit between the gates serves as a means of coupling the oscillator to the isolating output stage.

The upper/lower sideband switch applies +9 volts to the anode of the switching diode that closes the feedback circuit for the particular crystal to be activated. Since the unit is a transceiver, the output can be used as injection voltage in the demodulation of an incoming signal by the product detector or as the basic carrier applied to the balanced modulator for the transmit mode of operation.

Since the transceiver must operate over a wide frequency range the channel oscillator must be able to accommodate crystals over a 2- to 23-MHz range. The two dual-gate mosfets operate in the untuned crystal oscillator circuit in **fig. 18**. The Colpitts type crystal oscillator is followed by an isolating amplifier stage. A small netting capacitor can be used for netting an individual crystal to a precise assigned frequency.

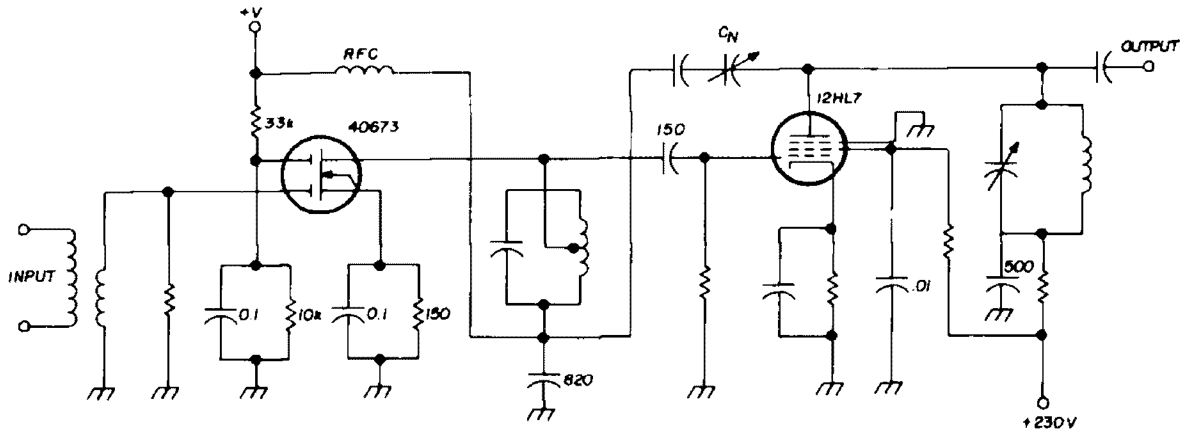


fig. 20. Using a mosfet amplifier to drive a vacuum-tube rf power amplifier.

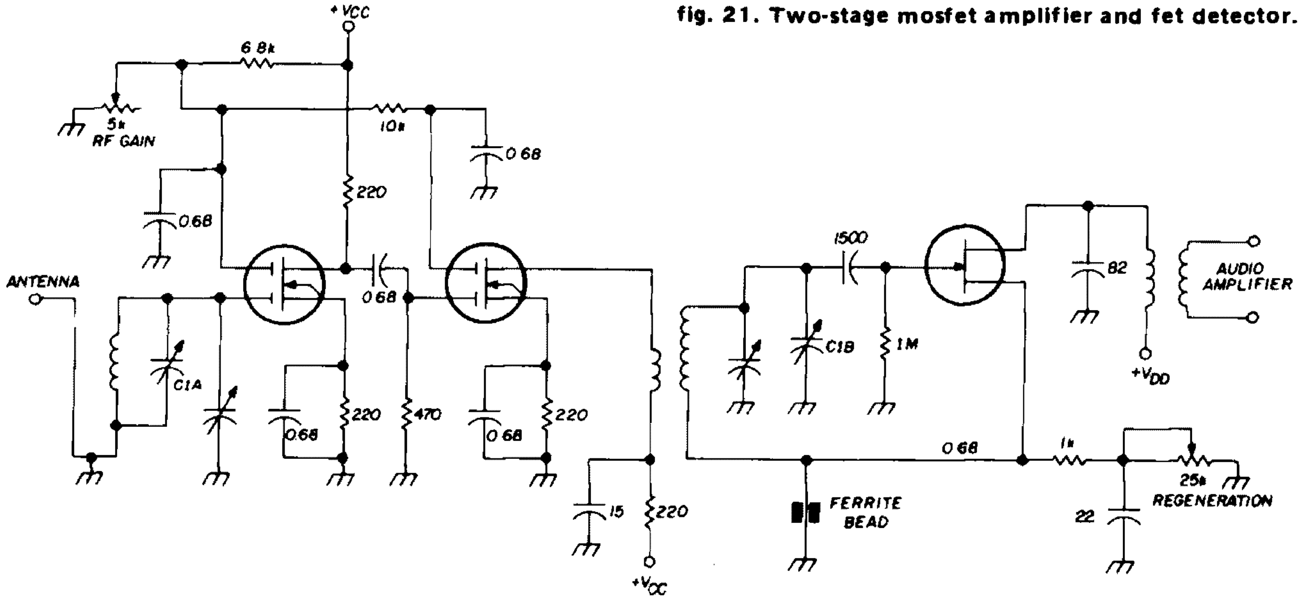


fig. 21. Two-stage mosfet amplifier and fet detector.

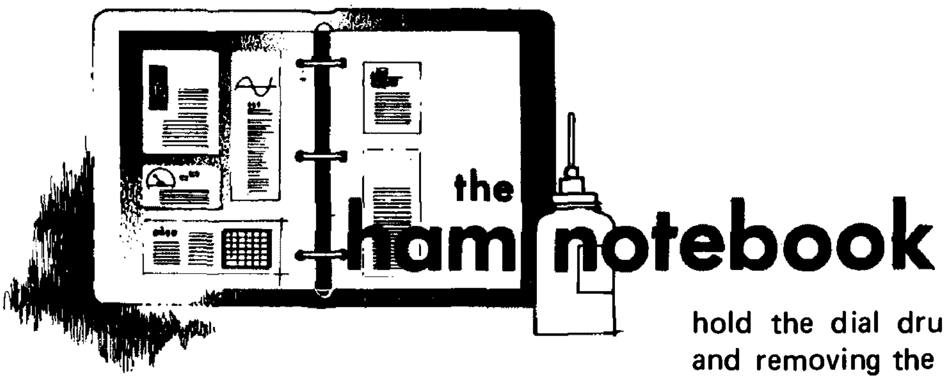
A transmit mixer is shown in fig. 19. The low-frequency sideband signal and high-frequency oscillator signal are mixed to produce a higher sum frequency at the output. A double-tuned resonant circuit provides adequate output bandwidth and excellent skirt rejection of undesired frequency components.

The ease with which a mosfet can be used to drive a vacuum-tube stage should be attractive to those of you who want to build a hybrid transmitter but just haven't gotten around to it. Fig. 20 shows a simple circuit arrangement that permits you to drive a modest power pentode with a mosfet. The channel frequency signal is applied to the input gate, amplified, and is resistor-capacitor coupled to the grid of the vacuum tube. To prevent instability and possible self-oscillation when operating over a wide frequency range a simple capacitive feedback link can be used. The neutralizing capacitor is adjusted for an optimum setting that covers the desired transmit frequency range.

If it's something different you wish to experiment with, and perhaps something that could be made quite effective for specific needs, take a look at the circuit of fig. 21. This is a two-stage mosfet rf amplifier followed by an fet regenerative detector (a TRF receiver). All I know about the circuit is given in the schematic. Maybe you'd like to expand upon it and come up with a small, low-power CW receiver. Two applications that come to mind are the 160-meter CW spectrum and the isolated segment of the 10-meter band assigned to novice operation. No doubt the receiver could be adapted to multiband operation as well as coverage over a wider frequency band with appropriate trimmers.

The gain control regulates the voltage applied to the second gate of both rf amplifiers. The dual-gate connection permits this isolation of gain control and signal circuits. The feedback path and regeneration control circuit is located in the emitter circuit of the fet detector.

ham radio



correcting mechanical backlash in the Collins 70K-2 PTO

When the 70K-2 PTO in my Collins 312B-5 remote vfo developed an annoying mechanical backlash, I decided to dig into it to see if I could fix it myself and save the $75 or so Collins currently charges for a rebuilt PTO. Initially, I was hesitant to work on the PTO for fear of upsetting the frequency calibration, but when the problem became sufficiently annoying that I had either to fix it or replace it, I decided I had nothing to lose by trying. As it turned out, the remedy was simple and did not require disturbing any of the internal wiring of the PTO. The frequency calibration was unaltered, and the backlash was completely eliminated. The 70K-2 PTO is also used in the KWM-2, 32S-1, 32S-3, 75S-1 and 75S-3.

The symptom of the problem was that in *certain areas* of the PTO range, I could rock the dial back and forth up to 1 kHz without changing frequency. Of course, mechanical backlash can also be caused by a malfunctioning dial mechanism, but in this case I determined that the problem was in the PTO itself. The first thing to do is to get the PTO out in the clear where it can be easily worked on by loosening the two set screws that hold the dial drum on the PTO shaft and removing the two mounting screws that hold the PTO rear cover to the chassis. With the 312B-5 there was sufficient slack in the wiring that it did not have to be disconnected, but this may be necessary on the other equipment. Once the PTO is dismounted from the chassis, the rear cover can be removed and slid down the wiring harness out of the way.

On either side of the PTO shaft there is a screw which extends the entire length of the tuning coil form and is threaded into the tuning coil rear cover. Turn the shaft back and forth through its entire range a few times and note how the heads of these screws work in conjunction with the special bushings and washers on the shaft to act as stops, limiting the total shaft rotation to 2¼ turns. The bushing and washers should not be removed or loosened, as they determine the end points of the tuning range. It is important to note their relationship with respect to each other and to the screw heads in order to properly reset the tuning end points when reassembling the PTO.

There is a single tiny ball bearing which is held between a dimple in the tuning coil rear cover and a dimple in the end of the tuning coil shaft. Holding the rear cover in place, remove the two screws and be careful not to lose the ball bearing when removing the rear cover.

Looking at the rear of the exposed coil tuning assembly, rotate the tuning shaft and observe how the coil core moves in and out, riding on a spiral groove cut into the shaft. The core assembly has an anti-backlash spring which rides on a groove on the inside of the form. There is a lubricant which appears to be a graphite compound used between the shaft and the core. In my unit there was an excessive build-up of this lubricant at certain places on the shaft and inside the core. This build-up was thick enough to overcome the tension of the anti-backlash spring, allowing the core to rotate instead of moving in and out, thereby causing the backlash. Turn the shaft until the core can be removed, and wipe the excess lubricant from the shaft and from inside the core. To reassemble the PTO, merely reverse the disassembly procedure.

John Becker, K9WEH

simple satellite antenna

K4GSX's article on simple stationary antennas for satellite use showed how a ground-plane antenna could be adapted to OSCAR use by slanting the driven element and using a matching section to accommodate the change of impedance.* I've been using a simple ten-meter inverted-vee antenna for OSCAR reception quite successfully since December, 1972. As can be seen in fig. 1, a tilted vertical is physically one-half of an inverted-vee or drooping doublet, but without the matching section or radials. Therefore, an inverted-vee is a double-tilted vertical fed in the center.

A dipole theoretically has a radiation resistance of approximately 72 ohms, and slanting the wires downward lowers the input impedance sufficiently for direct connection to 52-ohm cable. Other advantages *of the inverted-vee* include ease of construction and erection, and the possibility of adding a balun later, if desired.

The characteristics of an inverted-vee are somewhat different than those of a dipole, and it is suggested for satellite communications that the wires run north and south because the radiation pattern is off the ends of the antenna. This arrangement facilitates reception since the satellite travels in a polar orbit. Because of the antenna's physical characteristics, the inverted-vee probably offers the best compromise between horizontal and vertical polarization, and, therefore, is well suited for satellite work while still meeting the four design objectives cited in the article. Although no claims are made as to the radiation pattern, it should probably be close to that of a tilted vertical.

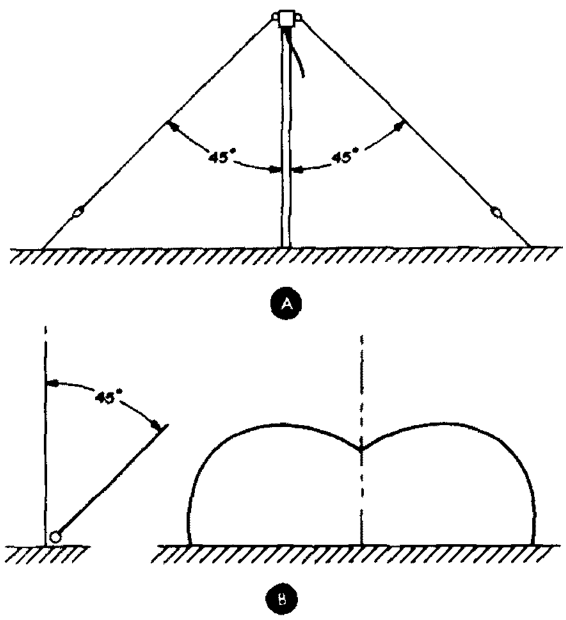


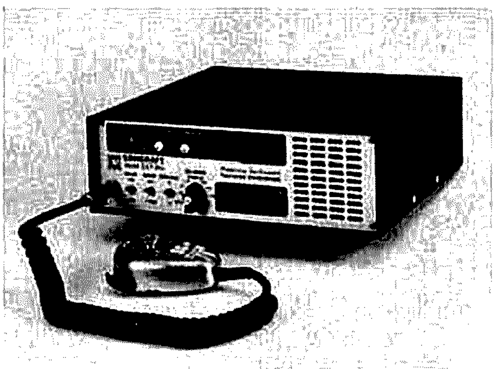
fig. 1. Diagram showing the physical characteristics of a quarter-wave tilted vertical, to that of an inverted vee. The inverted-vee in (A) uses two center fed quarter-waves. The tilted vertical in (B) in one quarter-wave, base fed.

Craig Caston, WA6PXY

*Dale Covington, K4GSX, "Simple Antennas for Satellite Communications," *ham radio*, May, 1974, page 24.

new products

two-band fm transceiver



The Comcraft Company has announced the introduction of a new all solid-state two-band, frequency-synthesized fm transceiver, the model CST-50. The CST-50 features operation on both two and 1-1/4 meters with 25-watts output and 5-kHz frequency-synthesized channel spacing. Operating modes provided include simplex, split transmit and receive, and repeater offsets of plus and minus 600 kHz, 1 MHz and 1.6 MHz.

Frequency coverage on the 220 band is from 220 to 225 MHz while on two meters it is from 142 to 149.995 MHz (to cover most MARS, CAP and CD frequencies). The frequency synthesizer is a digital type using programmable dividers and phase-locked loops to generate the desired frequencies by reference to a single 5-MHz crystal. Two thumbwheel sets are provided for setting up either transmit and receive frequencies for non-standard repeaters or two separate repeaters when standard offsets are in use.

Additional features of the CST-50 include the use of an eight-pole 10.7-MHz crystal filter in the receiver; transmitter coverage of both bands without retuning; illuminated thumbwheel switches for night mobile work; PTT mike and mobile mount; and quality construction using epoxy-glass PC boards and vinyl-clad outer cabinet.

The CST-50 is priced at $769.95. For more information, write to Comcraft Company, Post Office Box 266, Goleta, California 93017, or use *check-off* on page 94.

ARRL antenna book

The new 13th edition of *The ARRL Antenna Book* represents the most extensive revision this publication has received within the past 25 years. Although much of the basic information of previous editions on subjects such as radio propagation and antenna theory has been retained in early chapters of the book, all information has been carefully edited for clarity and has been supplemented with later data where modern technology has brought new knowledge.

In the later chapters some striking changes from previous editions will be noted. A large section appears on the use of the Smith chart in solving transmission-line problems. Information on cubical-quad antennas has been greatly expended. Design and construction information on log-periodic antennas has been added. Construction information on standard antennas — dipoles, Yagis and simple arrays — has been revised extensively, and new antenna types such as a 40-meter *sloper*

More Details? CHECK—OFF Page 94

are described. Information on rotator and tower selection and installation have also been added.

Four new chapters appear in the 13th edition, one on antennas for restricted space, one on antennas for space communications, one on measurements and one on specialized antennas that amateur radio enthusiasts often hear about but are unable to find information on — the Beverage, discone, conical monopole, fishbone, bobtail curtain and others. From its newly designed front cover, which retains a bit of the appearance of the covers of older editions, to its completely new index at the back, this edition is packed with useful information on all types of practical antennas.

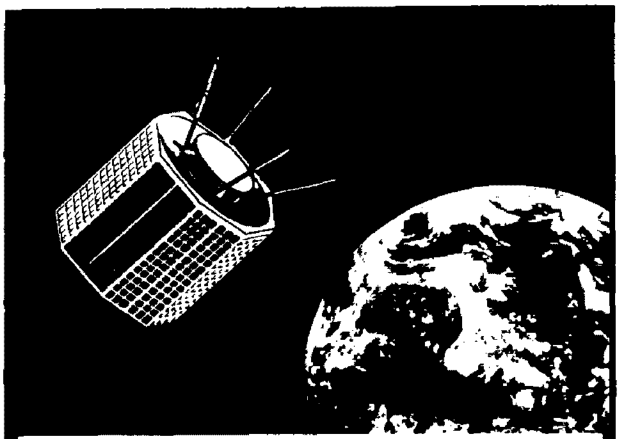
The new edition contains 336 pages and is priced at $3.00 from HR Books, Greenville, New Hampshire 03048.

160-meter transverter

The new 160-meter transverter from Dentron Radio provides up to 100-watts input over the entire 160-meter band. This new unit, which needs only two connections to your existing ssb transceiver, requires only five-watts drive for full rated output (3.8- to 4.0-MHz input). The transverter includes a built-in power supply for 115/230 Vac, 50/60 Hz.

Also available from Dentron is their new 160 AT antenna tuner which matches any 160-meter exciter with a 50-ohm output to almost any random-length wire or existing antenna.

The 160 XV 160-meter transverter is priced at $199.50, postpaid in the USA.



The 160 AT antenna tuner is priced at $59.95. For more information, write to Dentron Radio Company, 27587 Edgepark Drive, North Olmstead, Ohio 44070, or use *check-off* on page 94.

new signal/one

The new Signal/One CX-11 deluxe integrated station offers many new features not available on the older CX-7A. Featured in the new design is a broadband solid-state linear power amplifier with 175 watts rf output. It requires no tuning over the six amateur bands from 1.8 to 30 MHz, operates into any vswr and is capable of continuous duty at full rated output.

The new CX-11 also contains a new concept in front-end design — using doubly-balanced active fet mixers for unmatched sensitivity, blocking and cross-modulation rejection. Five bandwidths of audio selectivity are standards: 2.4, 1.5, 1.0, 0.4 and 0.1 kHz. A peak/notch filter with adjustable frequency notch depth is also included.

Also featured in the CX-11 is a built-in electronic keyer with independent speed and weight control and partial or full-dot memory. The six-digit frequency readout uses half-inch amber or red LEDs and is optimized for non-blinking, stable display. The power supply is completely self protecting — both thermal and current overload — and is IC controlled. Additional features include dual VFOs for transceive, split operation or dual receive, adjustable i-f shift, receive or transmit offset tuning, push-button spotting, adjustable rf clipping, instantaneous CW break-in, built-in wattmeter, built-in noise blanker and adjustable rf power output.

The Signal/One CX-11 is now in production at $2600, and is distributed by Payne Radio, Box 525, Springfield, Tennessee 37172. For more information, write to Signal/One, Box 127, Franklin Lakes, New Jersey 07417, or use *check-off* on page 94.

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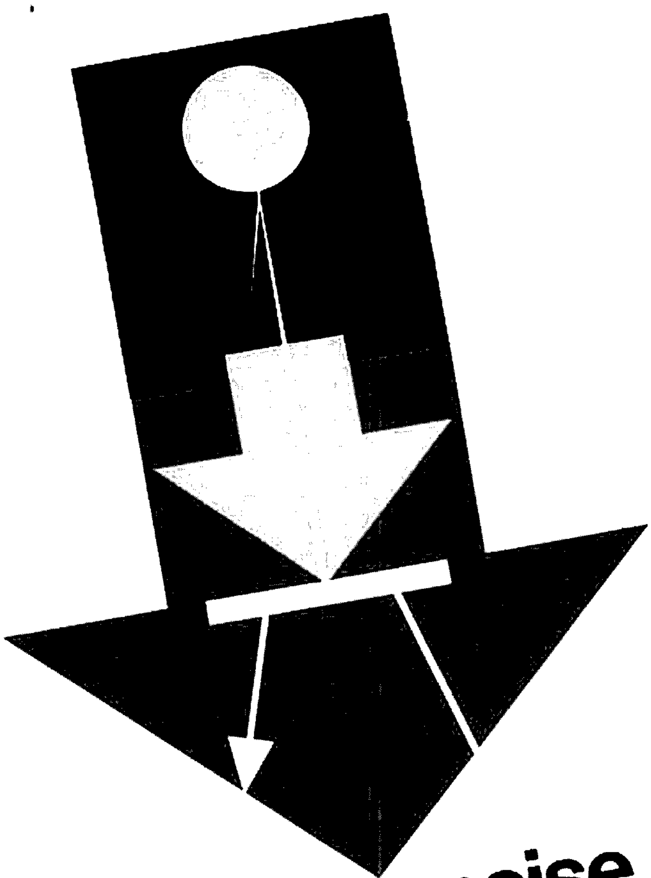

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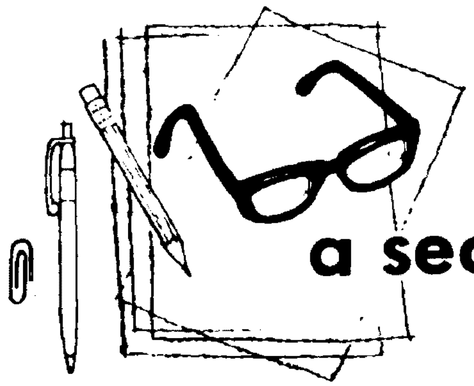
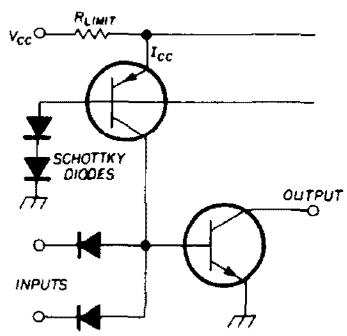


a second look

by jim fisk

Two of the major semiconductor manufacturers are working on new families of bipolar digital logic that may do as much to revolutionize the computer world as anything in the past. Texas Instruments, for example, has developed a new family of Schottky logic, called Schottky II, with 1-to-2-nanosecond delays (compared to 10 nanoseconds or so for TTL) which operates from a single 5-volt power supply, offers better performance than today's ECL 10k and is provided in standard, easy-to-use pinouts. Unfortunately, because of the dismal market conditions which are facing the semiconductor manufacturers just now, it may be months before Schottky II is available to designers.

Although not a great deal is known about Motorola's new bipolar logic family, it is known to feature 1-to-2-nanosecond delays and will be compatible with ECL 10k. However, they have been working with a system of complementary constant-current logic (C^3L) with 1-nanosecond delays (and 1 mW per gate) that could provide the necessary performance. Motorola is also rumored to be working on a family of sub-nanosecond logic called ECL 100k that should provide some answers for digital designers who are looking for faster and faster computers.



Nand gate built with new C^3L logic features speed, low-power consumption and high packaging density.

The complementary constant-current logic is particularly interesting because it combines the best of several worlds not previously available on one chip: very high speed, high packaging density and low power consumption. The high packing density of C^3L is due to its very simple transistor-gate structure, shown below, which consists only of a pnp current source transistor, an npn switching transistor and a few Schottky barrier diodes. By way of contrast, the Schottky TTL gate uses four transistors, plus diodes.

Because of the simple structure, a five-output gate requires only 12 square mils (0.008 square mm). By comparison, a low-power cmos circuit of the same complexity requires about 65 square mils (0.04 square mm) of chip space. This high packaging density means that a 1000-gate C^3L array could be placed on a 150-by-150-mil (4x4mm) chip.

While you may be hard put to think up an application in your amateur station for a 1000-gate logic array, many traditional analog circuits (TV tuners, for one) have gone the digital route in recent years, and other devices, such as frequency synthesizers, small hand-held calculators and digital meters, would never have seen the light of day in an all-analog world. If the past is any indication, future digital applications will have an even far wider effect.

Jim Fisk, W1DTY
editor-in-chief

presstop

AMATEUR LICENSE FEES REDUCED with reductions to go into effect March 1. New fee schedule is $4 for a new license or renewal, $3 for a modification. Requests for special calls will still cost $25.

WWV/WWVH PROPOSE CUTBACK in services, solicit advice from users. Write Time and Frequency Services, National Bureau of Standards, Boulder, Colorado 80302 for detailed questionnaire.

BICENTENNIAL CALLSIGNS pretty well set, will use only AA-AL prefix block to cover both continental and offshore W and K prefixes during 1976 celebration. Though stateside bicentennial prefixes will use present numbers to indicate call areas, look for some real odd balls for such things as Alaskan Novices...

220-MHZ CLASS-E CB still very much a threat -- recent letter from OTP Acting Director Eger to FCC Chairman Wiley urges giving "every consideration" to "expeditious action" in granting Class-E 222-224 MHz! The letter cites the need for a "disciplined" citizen's communications service, half-billion dollar a year market, gives lip service to value of amateur service -- and would let us continue to use the new CB band, with limitations! Class E could start up as early as May!

GOVERNMENT AGENCY SHAKES ANTENNA TOWERS -- OSHA, the occupational safety people, want all towers over 20 feet to have a built-in OSHA-approved ladder (16-18" wide, with side rails mounted 6" off the tower)! Since the OSHA ladder is heavier than most light-duty amateur and TV antenna towers, such a requirement would have a great effect on tower prices. Ruling would become effective this summer barring protests, but look for lots of those from users and makers alike.

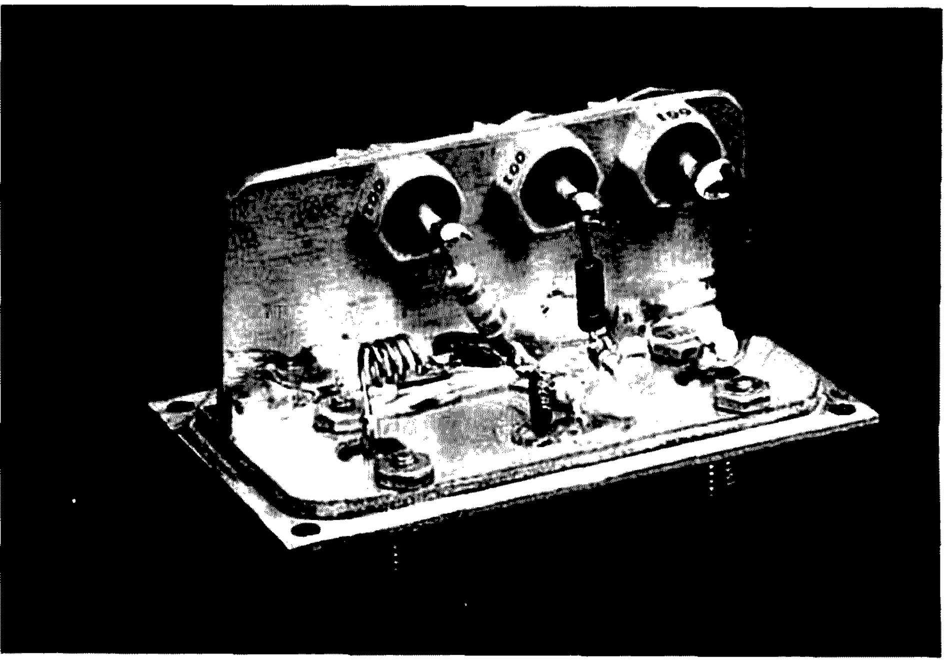
ARRL DIRECTORS' MEETING in January confirmed Dick Baldwin, W1RU, as League General Manager and QST editor to replace John Huntoon. Dick took over February 1, while John continues as both ARRL and IARU Secretary.

Docket 20282 Discussion occupied much of the Directors' time, resulted in authorization of a "membership opinion survey" to be conducted on the Division level. Survey will be made in time for individual directors to consider results before a special board meeting in May when League response will be determined.

NATIONAL ENVIRONMENTAL POLICY ACT OF 1969 appears likely to have little direct impact on amateur radio, but indirect effect on the unwary may cause lots of bother. Environmental impact report must be filed with amateur license applications on Form 610 or 610B only if the station will have a "major" impact on the environment. Installations having a "major" impact are: Antenna towers or antennas over 300 feet high (but not an antenna mounted on an existing structure 300 or more feet high); satellite earth stations with dishes of 30 feet or more in diameter; locations in a wilderness area, wildlife preserve or national scenic or recreation area; stations affecting sites significant in U.S. history; and any involving extensive changes in land-surface features.

Applications Other Than The Above are considered "minor" and no statement need be filed. However, to be on the safe side, until new forms are available, add a statement such as, "This application is a 'minor action' as defined by Section 1.1305 of the Commission's rules." somewhere on the form unless you will have a "major" environmental impact, in which case a call to the FCC is advisable.

NEWLY ANNOUNCED HAND-HELD CALCULATORS, as promised in January editorial, offer more performance for less cost. New HP-21 Scientific from Hewlett-Packard ($125) is more powerful than HP-35 and features 32 pre-programmed functions including rectangular/polar conversion, while new programmable HP-55 ($395) provides 49-step programming and 86 keyboard functions. Also new are Novus Mathematician ($80) and 10-digit Corvus 500 ($200).



ultra low-noise uhf preamplifier

Design and construction of an ultra low-noise preamplifier with a 1.0 dB noise figure at 432 MHz that provides high performance on 144 and 220 MHz as well

Joseph H. Reisert, Jr., W1JAA, Chelmsford, Massachusetts 01801

There is always a need for a better preamplifier with a lower noise figure and higher gain. Such a preamplifier was introduced in November, 1972.[1] This preamplifier took up the slack after the TIXM05 disappeared and the low-noise fets bottomed out, and it introduced several new features to amateur radio including state-of-the-art noise figure, wide bandwidth, low-Q input, current-source biasing and built-in limiter. In addition, it required no tuning.

Since the original preamplifier was introduced, an improved version has been developed. This new preamplifier has higher gain, lower noise figure, and an improved biasing scheme while embodying all the other features mentioned above. It has been duplicated by over twenty-five individuals throughout the world and is the input preamplifier used at most of the 432-MHz EME stations. Two of the die-hard paramp users now have models of these preamplifiers in use. It meets or exceeds their paramp performance and is easily

mounted at the antenna, a feature not easily duplicated with paramps and their associated pumps.

requirements for low noise figure

A low-noise-figure transistor is required in the circuit but it is not the only requirement for a low-noise-figure preamplifier. Other requirements include proper operating current and voltage, optimum source impedance, low-loss matching circuits, low feedback, moderate gain and good stability, to mention a few. Let's discuss these requirements separately.

The need for a low-noise-figure transistor should be obvious. You cannot attain a noise figure which is lower than the device is capable of delivering due to other factors affecting the design. You will be lucky, at best, if you end up within 0.25 to 0.5 dB of the device's capability.

Joe Reisert, W1JAA, was first licensed in 1951 as WN2HQL, and in 1956 earned his Extra Class license. He moved from Long Island to San Jose, California, in 1961, where he was licensed as WA6TGY, later as W6FZJ. He attained the DXCC Honor Roll in 1968 and presently stands at 330 confirmed. In the late 1960s he became interested in uhf, had his first 432-MHz contact in 1970, and put his EME station on the air in 1972. Before moving to Massachusetts last spring he had worked nine states on 432 MHz from California plus Canada and Australia. He is joint holder of the 2304-MHz tropo DX record of 330 miles set in February, 1974, and is active from Massachusetts on 432-MHz EME. Primary amateur interests are DX, EME, and antenna and receiver design.

Joe was formerly the supervisor of Microwave Product Engineering at Fairchild Microwave after working at Sperry, IBM, Lockheed and Wescom Microwave, and is presently manager of Microwave Applications Engineering at Alpha Industries, Inc., in Woburn, Massachusetts — a leading manufacturer of microwave diodes.

The collector current (I_C) and collector-to-emitter voltage (V_{CE}) are also prime considerations. Generally, V_{CE} is not too important if it is greater than 6 volts. Lower V_{CE} generally lowers the collector cutoff frequency (f_T) and hence, the gain. The collector current, on the other hand, is very important. Older devices were usually optimum with 1.0 to 1.5 mA collector current. The newer devices, as a rule, work best at 2 to 3 mA collector current and their noise figures do not degrade as fast as their predecessors' at higher collector currents (more on this later).

The optimum source impedance is the impedance that the transistor wants to see in order to deliver the lowest possible noise figure. At frequencies below 1000 MHz this value is seldom 50 ohms. Therefore, a matching network is usually necessary between the antenna input and the transistor. This network must have very low loss since any losses will add directly to the overall noise figure of the preamplifier.

In addition, low feedback is essential to low-noise operation and feedback will usually raise the noise figure. This also applies to series feedback in the emitter lead. The emitter should be well bypassed to ground to prevent degeneration and higher noise figures. Current-source biasing is preferred since it allows the emitter to be grounded directly without bypassing. A suitable scheme was used in the original preamplifier. This design will include a simpler and less critical circuit.

High gain is essential for low noise performance. If the gain of the preamplifier is not high enough, the overall system noise figure will be degraded due to the noise figure contribution from the second stage. However, if the preamplifier gain is too high it may become unstable or may overdrive the second stage and cause desensitization or intermodulation distortion. A good rule of

thumb is to strive for 10 to 13 dB minimum preamplifier gain. Gain above 18 to 20 dB should be avoided since it generally indicates a potential instability. More on this later. A method for computing gain and overall noise figure is discussed further in **appendix** 1.

tradeoffs

Noise figures below 1.5 to 2 dB are usually wasted on meteor scatter and tropo communications. At 144 MHz this is still a low enough noise figure for EME since the sky temperature is usually not much lower than the terrestial temperature. Therefore, a good fet is sufficient to deliver all the performance needed. However, on EME above 420 MHz the sky temperature may be below 70° K, so a lower noise figure is desirable. For example, a 1 dB noise figure preamplifier may yield up to 2 dB better signal-to-noise ratio than a 1.5 dB noise figure unit.

Cost is obviously the most important tradeoff. Low-noise-figure, high performance transistors are expensive. However, in view of the performance gained, it is penny wise and pound foolish to cut corners too closely. Generally speaking, a low-noise transistor suitable for tropo operation up to 450 MHz will cost about $10, and up to $25 for a higher performer. The sky is the limit when it comes to the extremely low noise figures (1.5 dB or less) necessary for EME.

In the preamplifier presented here a device costing less than $50 will outperform just about any device presently available *at any price*. On EME you may spend hundreds, and maybe thousands,

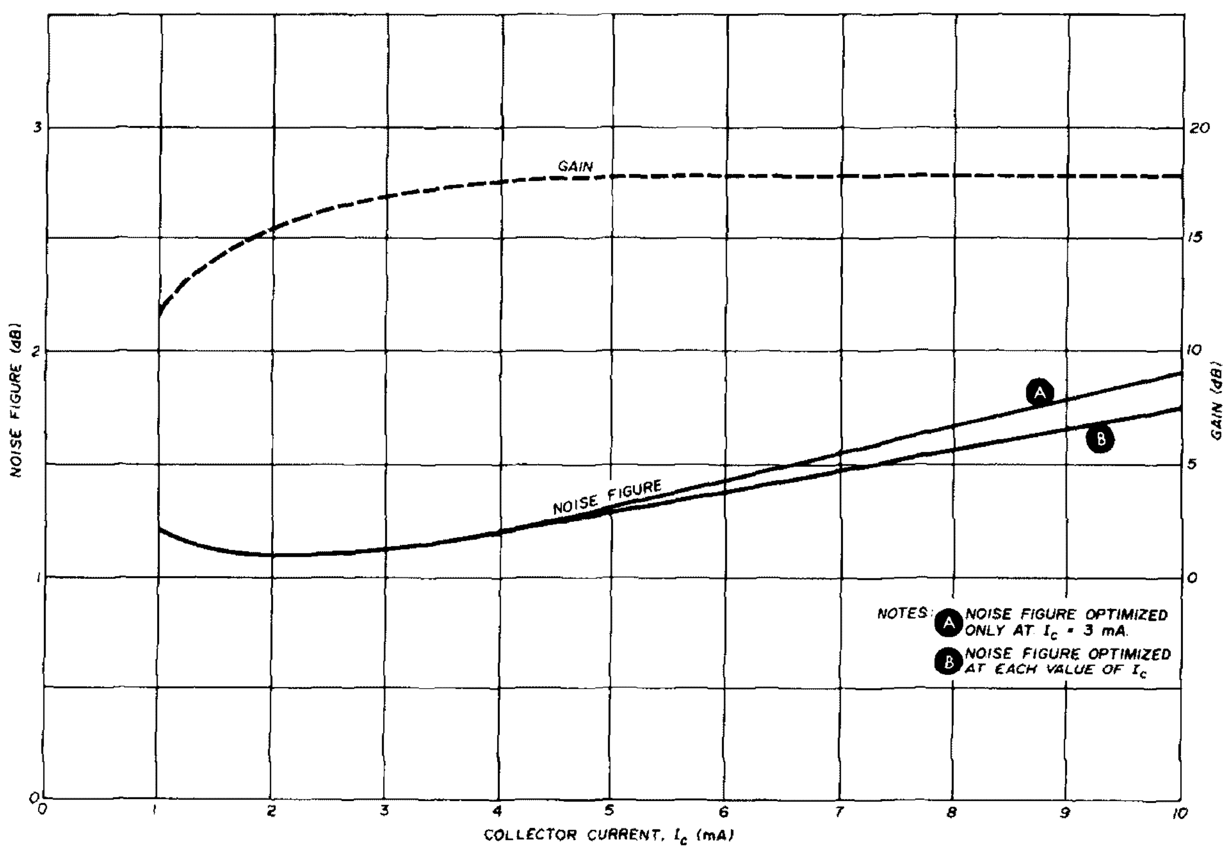


fig. 1. Typical FMT4575 noise figure and gain at 432 MHz vs collector current (V_{CE} constant at 7 volts).

of dollars and hours building a suitable station. Why skimp when it comes to the preamplifier? A suitable performance increase in the antenna may be completely out of the question due to size or money.

Gain isn't everything, but it does help. This is especially true when the preamplifier is remotely mounted. An extra gain margin helps to overcome cable losses. Furthermore, on the more modern low-noise transistors the gain increases quite smoothly when the collector current is increased while at the same time the noise figure may not rise as rapidly as you would expect (see fig. 1). As an additional by-product, the intermodulation distortion decreases significantly with increasing I_C. Again, it pays to use a high performance transistor so you can "have your cake and eat it too."

A broadband preamplifier can present some problems since there is little or no discrimination to out-of-band signals. Therefore, an input filter is highly recommended. A suitable type will be discussed in the latter part of this article.

transistor selection

I have evaluated many npn transistors, all with an eye on minimum noise figure and maximum stable gain at 432 MHz. The circuit described in reference 1 used the NEC 2N5650 series (and its offshoots such as the NEC V766). However, a 1.4 dB noise figure was the lowest measured, and stable gain seldom exceeded 13 dB. Other high performance devices which were tried included the Fairchild FMT4225 and Hewlett-Packard HP-21 series. They delivered high gain but higher noise figures (1.5 to 2 dB typical). Likewise, the Amperex BFR90 and BFR91 seldom provided noise figures below 2 dB.

The Fairchild FMT4575/4578 and FMT4000/4005 were true eye-openers; they easily yielded a 1.25 dB noise figure and 1 to 1.1 dB was not uncommon. Since the FMT4000/4005 and 4578 are higher priced and didn't provide better performance, I decided to design a preamplifier around the lower priced FMT4575. At $44 (each) it delivers the most performance per dollar of any transistor presently available and will challenge the best of paramps at 432 MHz.

For those interested in evaluating their own devices a lab test setup can be built as shown in fig. 2. The device to be tested is mounted in a suitable low-loss transistor mounting fixture which includes a 50-ohm input and output line with dc blocks. This mounting fixture is connected between two low-loss double- or triple-stub tuners. A noise-figure generator is connected to the input stub tuner and a very low-noise-figure converter is connected to the output tuner. Then the bias voltage and current are adjusted to predetermined values (per manufacturer's data sheets).

The output tuner is first adjusted for maximum stable gain and then the input tuner is adjusted for minimum noise

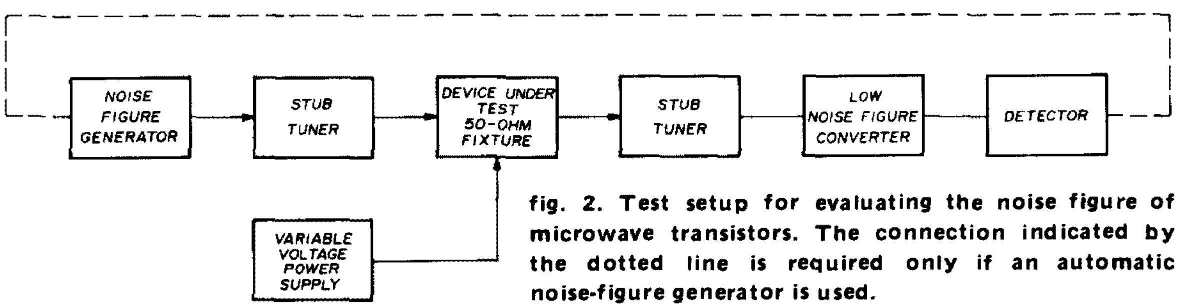


fig. 2. Test setup for evaluating the noise figure of microwave transistors. The connection indicated by the dotted line is required only if an automatic noise-figure generator is used.

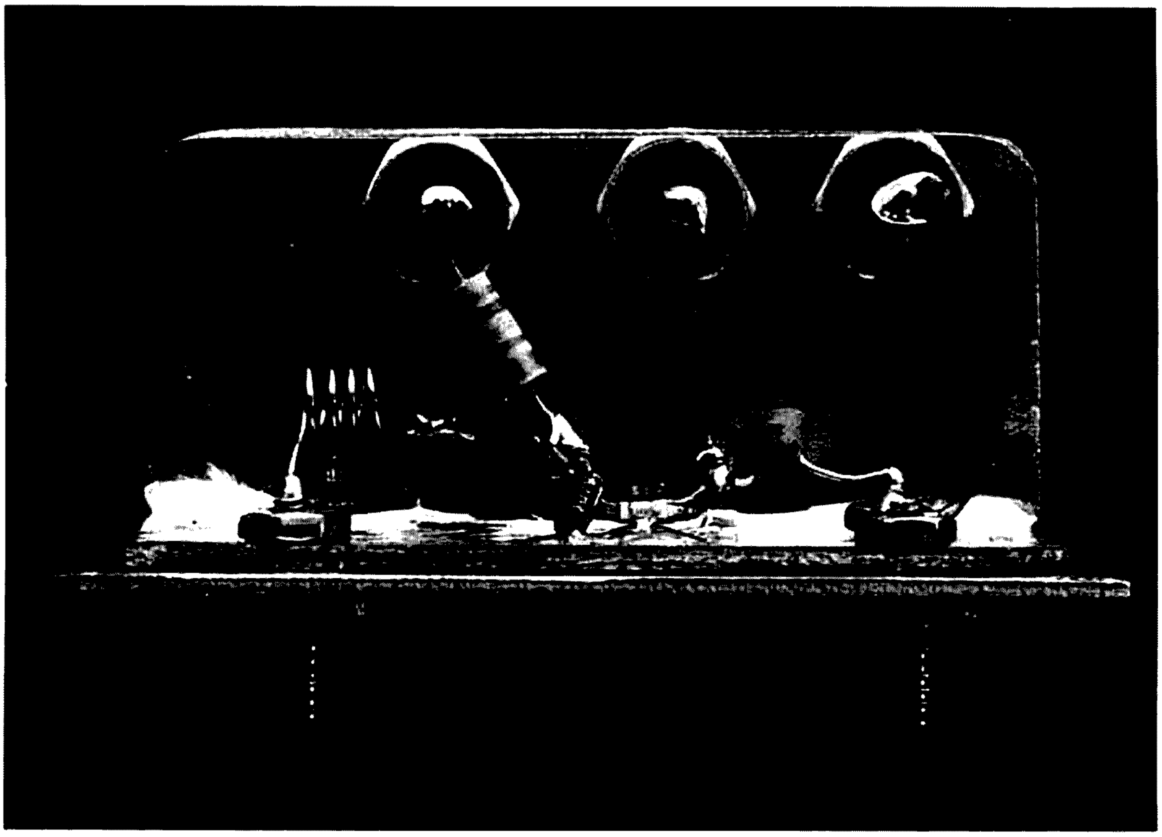
figure. After optimization the bias points are varied up or down (within manufacturer's ratings) and the tuners are readjusted. This procedure is continued until the lowest possible noise figure is obtained.

circuit description

While the preamp described here is quite similar to the original design, there are important differences. They are basically the transistor used, the input/output matching and the current source.

The transistor was chosen by the method described above. Then the transistor test fixture was split apart and a low-loss 50-ohm termination was substituted for the noise-figure generator. With the aid of a network analyzer, the desired source impedance was measured. This impedance is called the optimum source impedance as described earlier. In the case of the Fairchild FMT4575, this impedance can be approximated by a series circuit consisting of a 50-ohm resistance and an inductive reactance of 80 ohms (see fig. 3).

It so happens that at 432 MHz most of the low-noise transistors evaluated generally required a similar network except that the values varied from 25 to 75 ohms for the resistance and from 30 to 100 ohms for the inductive reactance. These values can be easily simulated by slight changes in the inductance value and by placing a small (0.3 to 3 pF) low-loss variable capacitor either between point X or Y to ground (fig. 3). However, this capacitance value is not critical and seldom yielded much improvement in noise figure on the devices I tested.



Construction of the ultra low-noise preamp showing placement of the main components. Components can be identified from fig. 7 on page 17.

A low-loss input matching circuit is very important. Therefore, low-loss components should always be used with the least complicated, low-Q circuit.

This is in direct contrast with previous design philosophy which frequently used filters and/or resonant input circuits. Such circuits can contribute additional losses. *The input of a low-noise preamplifier is a poor place to obtain selectivity*. A better choice is to install a low-loss filter external to the preamplifier as discussed later.

The final input matching circuit chosen was a low-loss, low-Q, L-matching section consisting of L1 and CR1 (**fig. 4**). Capacitor C1 is a blocking capacitor (not a critical value). However, a low-loss, high-Q type is desired. RFC1 is essentially a low-Q parallel-resonant circuit at 432 MHz and therefore is effectively out of the circuit. It actually works quite well from 100 to 450 MHz. Some of the physically larger RFCs available are parallel resonant below 450 MHz and are not recommended. A grid-dip oscillator can be used for a quick test or, you can wind your own choke using a 0.1 to 0.2-inch (2.5 to 5.0 mm) diameter air core. A higher inductance RFC can be used if only lower frequency operation is desired.

Do not leave out the hot-carrier diode, CR1. It is the capacitance part of the L-matching section and adds about 0.75 pF to the circuit. Other hot-carrier diodes can be substituted provided the capacitance is 0.5 to 1.0 pF at zero volt. Do not use ordinary silicon or germanium diodes since they may increase the noise figure.

Diode CR1 also functions as a low-loss limiter and can save the transistor from being damaged if the preamplifier is subjected to excessive rf. Even rf from a high-frequency transmitter operating near a vhf antenna can do damage. This type of limiter is simple and effective. An added advantage is that it is placed after the selectivity. Hence, it will only activate when a strong input signal is present — it *will* not generate extraneous signals such as is common with back-to-back diodes connected across the input of a preamplifier ahead of the selectivity.

The bias scheme is a modification of the current source used in the previous design and is a variation of a method proposed by Fairchild Semiconductor.[2] I refer to it as "zener-diode biasing." It is much simpler than the transistor current-source and is less prone to oscillate. This bias scheme allows the

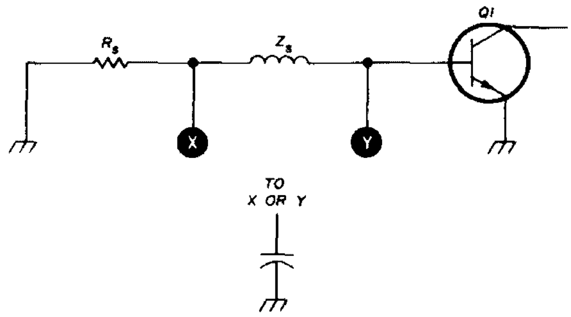


fig. 3. Rf equivalent circuit for an optimum source impedance network as described in the text.

emitter to be grounded directly, is insensitive to transistor current gain, provides some dc protection to the transistor and requires no adjustments.

Fig. 5 is a simplified circuit of the zener-diode biasing scheme. The zener diode, CR2, sets the transistor collector-to-base voltage (V_{CB}), R3 sets the I_C with a fixed supply voltage and R1 provides a keep-alive current flow for CR2. CR2 also provides protection to Q1 and limits the collector voltage to a fixed value. Once the proper values are chosen the transistor can be changed without any re-adjustment. The operation of this circuit is described in detail in **appendix 2**. CR3 (**fig. 4**) is an *idiot* diode (if you leave it out you're an idiot).

The output matching scheme is simplicity in action. This transistor (and many others like it) is so "hot" that all attempts at output matching *caused* instabilities. A computer program called SPEEDY[3] was called into action and a

program was written to select an output network which was unconditionally stable (will not oscillate regardless of input or output load). The final network turned out to be a 37-ohm collector resistor without any other matching elements. However, this lowered the gain too much. It was determined empirically that the collector resistor R2 can be raised to 100 ohms and seldom causes any instabilities. Therefore, if the input to your preamplifier and converter is highly reactive (most are), then the 100-ohm resistor may have to be lowered accordingly. This type of loading is also applicable to other preamplifiers which are only conditionally-stable.

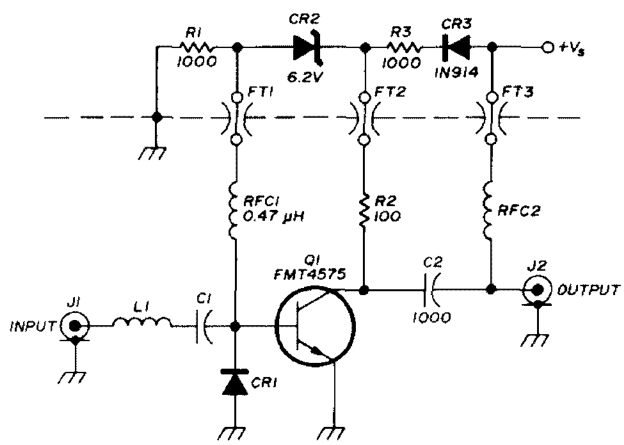
Finally, a simple "bias-tee" consisting of RFC2 in the preamplifier and RFC1 and C1 in the bias tee itself (see **fig. 6**) was added in case remote installation is desired. It not, this circuitry can be eliminated.

construction

Fig. 7 shows the preamp construction and the FMT4575 pin configuration. A small cast aluminum box such as the Pomona 2417 (2.25x1.375x1.125 inches or 57x35x29mm) is recommended. This box is available from most major electronic suppliers for $1.60. However, almost any shielded box is acceptable. The entire preamp is built on a small 1-1/8x2-inch (29x51mm) piece of double-clad glass-epoxy printed-circuit board which is attached to the box cover by the connector screws. If remote installation is not desired, FT3 can be mounted through the PC board and cover for power supply connections. This method of construction is versatile and simplifies soldering and testing. If you use the Pomona box, *be sure to remove the paint where the lid contacts the box. Also file off the anodized coating on the edge of the cover which contacts the box*. Failure to do so may result in erratic operation.

Be sure to check the data sheet for proper connection of any transistor other than the FMT4575. There is no industry standardization for lead identification on microwave transistors. The emitter lead length on this type of microwave package is not too critical and 1/8 inch (3mm) is recommended for operation up through 500 MHz. Both emitter leads should be the same length and both should be grounded. The leads of CR1 should be kept as short as possible so that it performs like a capacitor.

The rf connectors should be a good



C1 100-180 pF miniature dipped mica or ceramic

C2 1000 pF miniature ceramic disc CR1

CR1 hot-carrier diode (Hewlett-Packard 5082-2810)

CR2 6.2 volt zener diode (1N4735)

CR3 silicon diode (1N914)

FT1-FT3 feedthrough capacitors, 470-1000 pF

J1,J2 SMA-type coaxial connectors (see text)

L1 4 turns no. 24 on 0.1" (2.5mm) diameter, spaced wire diameter (approximately 30 nH)

Q1 Fairchild FMT 4575 low-noise transistor (see text)

R2 100 ohms, 1/4 watt (see text)

RFC1 0.47 µH miniature rf choke (Nytronics SWD=0.47)

RFC2 0.2-0.47 µH miniature rf choke or Ohmite Z-460 (value not critical)

fig. 4. Schematic diagram for the ultra low noise 432-MHz preamplifier. RFC2 is required only if the preamplifier is installed remotely.

uhf type such as SMA (OSM®). Discontinuities in BNC connectors can cause noticeable noise figure increases when operating with such a low-noise-figure device. Type-uhf connectors are definitely unacceptable at 432 MHz. Type N or TNC are also acceptable, but may be too large if a small box is used. An inexpensive version of the SMA connector is manufactured by the E.F. Johnson Company (JCM series, part number 142-0296-001) and is priced under $2.00. The type of output connector is not critical.

operation and test

After the preamplifier is assembled, a careful check of the wiring is recommended. Next, the preamplifier should be connected to a +12 volt power supply through a milliammeter (0 to 10 mA recommended). Terminate the preamplifier with a 50-ohm input and output load if available. If not, connect an antenna and converter to the preamplifier.

For proper operation, the total current should be 3.5 to 5.0 mA. If the indicated current is greater than 5 mA, remove power and recheck circuit wiring. High current usually means either a short circuit and/or improperly connected zener diode. If the power supply is variable, bring the voltage up slowly. At +11 volts, the current will be about 1 mA less than at +12 volts; with a +13 volt supply the current will be about 1 mA higher than at +12 volts. This indicates proper operation of the zener diode biasing circuits.

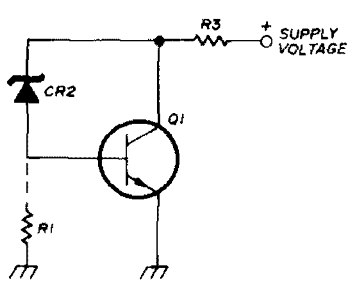
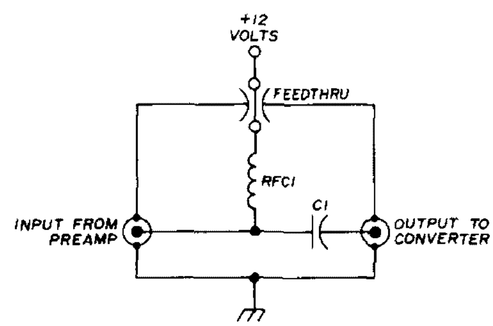


fig. 5. Dc schematic of the zener-diode bias scheme. Typical values are given in appendix 2.

It should not be necessary to make any adjustments. However, if you have sufficient test gear available, adjust the turns spacing on L1 for best noise figure by pulling apart or squeezing. This usually will only vary the noise figure by ±0.1 dB. For those purists who want to get all they can squeeze out of the circuit, an additional 0.3 to 3 pF low-loss piston capacitor can be placed between the input or transistor side of L1 to ground (see figs. 3 and 4).

input filter

It is advisable to use a low-loss,



C1	100-200 pF small dipped mica or ceramic capacitor
FT1	470-1000 pF feedthrough capacitor
RFC1	0.3-0.47 μH miniature rf choke (Ohmite Z460)

fig. 6. 432-MHz bias tee for use when the preamplifier is installed remotely is built into a small shielded box. Be sure to keep the leads on capacitor C1 as short as possible.

quarter-wavelength cavity filter ahead of the preamplifier since the simple broadband input circuit may cause intermodulation products from out-of-band signals. A suitable filter is shown on fig. 8. It should be connected as close to the preamplifier as possible. This can usually be accomplished with a short coax adapter. Multiple-element filters such as comb-line and interdigital types are not recommended. The input vswr to this, and most other, low-noise preamplifiers is typically 5:1. If a multiple-pole filter is used, it may suffer severe passband ripple due to the high vswr. The net result may be additional loss

and, hence, higher noise figure at the operating frequency. It may also cause out-of-band oscillations.

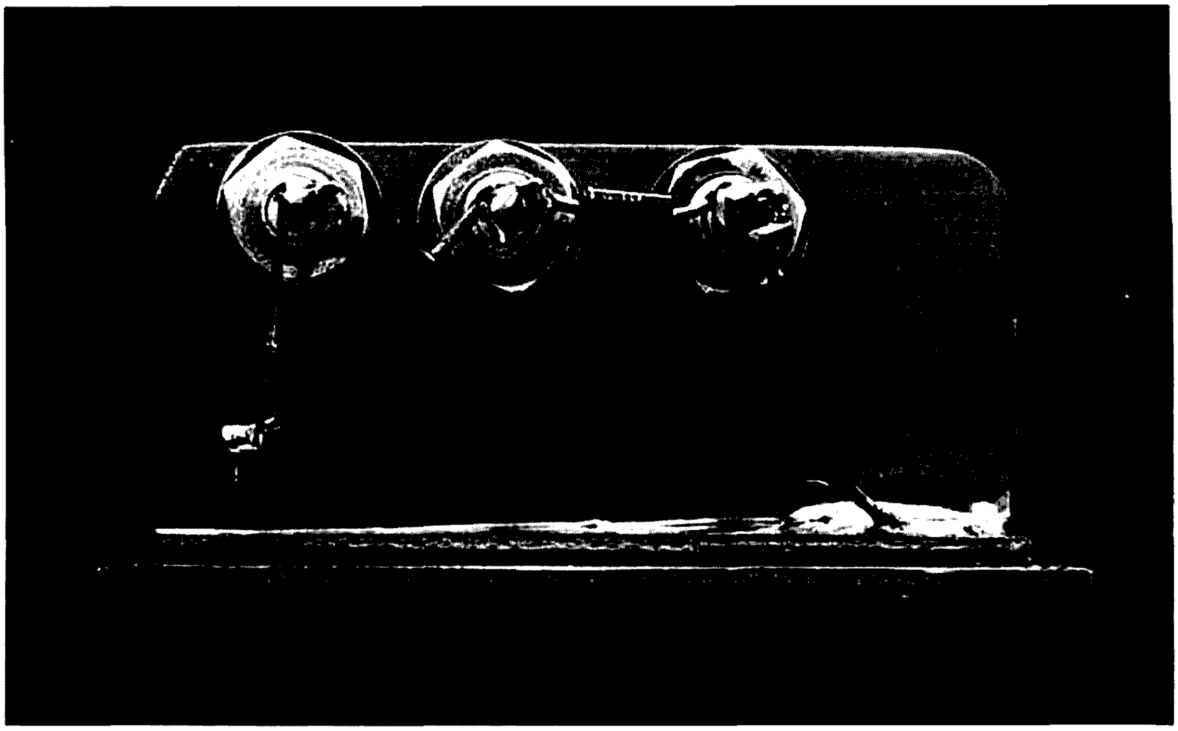
A word about the use of the proposed filter may be in order. After many hours of testing and re-testing various quarter-wavelength filters, I have concluded that a filter should not be tuned while connected to the preamplifier. Best results occur when the filter is adjusted by itself for minimum vswr with a good 50-ohm termination. Then the filter adjustment is locked in place. No further adjustments should be attempted if best results are to be achieved. Just connect the tuned filter to the preamplifier and accept its performance.

If very sensitive test equipment is available, the overall stability of the preamplifier may be tested. The reverse loss (inverting the input and output connections) should be at least 8 to 10 dB greater than the forward gain. The greater this margin, the less chance that the preamplifier is only conditionally stable. Many uhf preamplifiers I have tested only showed 2 to 4 dB difference, a clue to their instability.

performance

The ultra low-noise preamplifier really holds up to its title. The noise figure, when measured with a low-noise converter (2 dB maximum noise figure), is typically 1.2 dB. Some of the FTM4575s have even been below 1.0 dB and few have ever gone above 1.25 dB. The collector current can be easily adjusted by varying the supply voltage ±1 volt. The lowest noise figure usually occurs at 2 mA I_C. With the circuit shown in fig. 4, this will measure 3 mA to the preamp (subtract 1 mA, the keep-alive current of the zener diode — see appendix 2).

A typical plot of noise figure and gain versus collector current is shown in



Rear view of preamplifier showing locations of R1, R3, CR2 and CR3. Components can be identified from fig. 7 on next page.

fig. 1. However, at 2 mA I_C, the gain will be 1 to 1.5 dB lower and the intermodulation performance will drop by 6 to 8 dB. Therefore, 3 to 3.5 mA I_C is recommended and is set by the values in fig. 4. It is interesting to note that the noise figure is optimum from 1.0 to 5.0 mA I_C and therefore no re-tuning is really necessary as only slight improvement is possible at higher collector currents.

If a variable power supply is available, the operating point can be smoothly and easily varied up to 10 mA I_C for improved gain and intermodulation performance. The noise figure will typically only be degraded up to 1 dB under these conditions; if only a fixed supply is available, R3 (see fig. 4) can be changed to 390-ohms and an external 1000-ohm potentiometer (wired as a rheostat) can be used to vary I_C. In this case a milliammeter in series with the preamp power supply input is recommended. At extremely cold temperatures the noise figure decreases while the gain increases. Hence, instabilities can occur; they are easily rectified if I_C is varied accordingly.

noise-figure measurements

One last point may be in order concerning noise-figure measurements. Below 500 MHz there are considerable discrepancies when using automatic noise-figure measuring gear and units using the 5722 noise tube. Above 30 MHz these devices tend to generate additional excess noise which tends to make noise figures *look better than they really are*.

For instance, at 30 MHz the output of a typical automatic noise figure generator is 5.2 dB excess noise while at 432 MHz the excess noise can climb to 6.4 dB. If this is not accounted for the preamp may show a 1.2 dB better noise figure than the true value. This discrepancy is well known by the manufacturers and typical values are available. However, manufacturers are reluctant to compensate the older models since it would invalidate all test data and units in the field.

On the Hewlett-Packard 343A this effect can be easily compensated for by lowering the diode current from 3.31 to 2.5 mA at 432 MHz. Errors on the Hewlett-Packard 349A gas tube are less, typically only 0.4 dB. Newer units such as the AIL-75 should not have this problem. The so-called "hot-cold" method of testing is the most accurate system but is tedious and requires liquid nitrogen.

The noise figures quoted in this article have been tested by the hot-cold test method and with automatic test gear that was compensated. Therefore, the results are true noise figures, not ad-

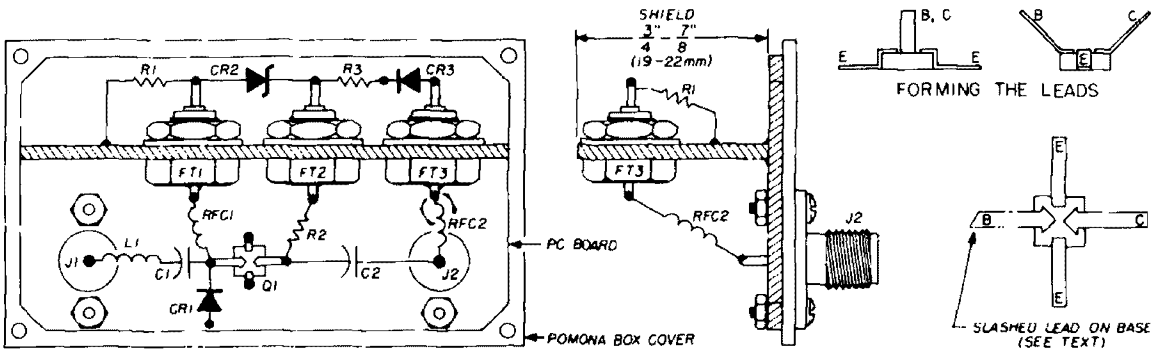
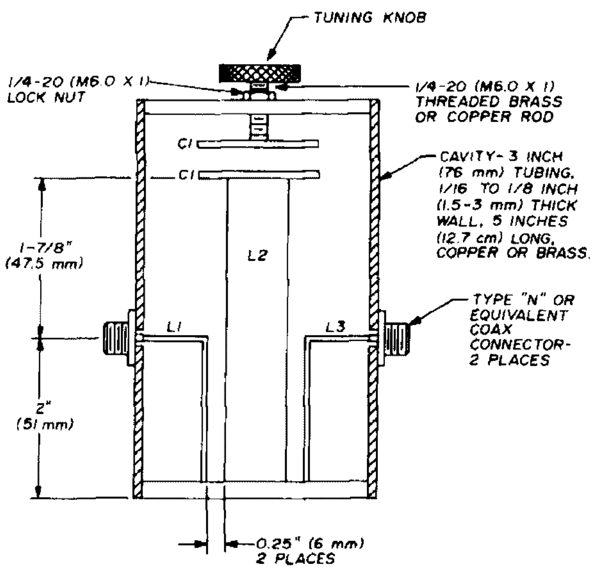


fig. 7. The ultra low noise 432-MHz preamplifier is built on a small section of double-clad printed-circuit board which is attached to the cover of the cast aluminum box.

vertising propaganda. Results have been widely correlated at the National Bureau of Standards, Boulder, Colorado; CSIRO (Commonwealth Scientific and Industrial Research Organization), Sydney, Australia, and elsewhere. In most cases this preamplifier exhibited 1.0 to 2.0 dB lower noise figures than the once revered low-noise standards such as the TIXM05, AF239, etcetera.



C1 1.5-2.0 inch (38-51mm) diameter brass discs soldered to tuning shaft and L2

L1,L3 number-12 copper wire spaced 0.25 inch (6.5mm) from L2

L2 0.75-0.875 inch (19-22mm) OD copper or brass tubing

fig. 8. Typical quarter-wavelength cavity filter for use on 432 MHz. Loss of this design is less than 0.2 dB and 3-dB bandwidth is 15 MHz (typical). Sweat solder the lower ends of L1, L2 and L3 to the base plate. Adjust for minimum vswr with 50-ohm termination and lock tuning control. Cavity may be silver plated for long-term, low-loss performance.

other variations

As pointed out earlier, this preamplifier works well at other frequencies although performance deteriorates above 500 MHz. The circuit was not optimized for other frequencies. However, even as is, the noise figure will be less than 2 dB at 144 and 220 MHz. Some slight adjustments will easily optimize operation at any frequency from 50 to 1000 MHz.

Other transistors will probably work well in this circuit "as is." However, gain may be lower, the stability poorer, and the noise figure higher. Therefore, some adjustments or changes may be necessary, but these changes have already been discussed and should not be a problem.

summary

This preamplifier should bring you right up to the state of the art in noise figure. It is simple to build and operate with no adjustments necessary. Don't forget the input filter, especially if you operate in a strong rf environment. Remember that this problem may be worse when the preamp is antenna mounted since the input line losses are usually lower.

It is hoped that this article will inspire more vhf/uhf building and operating activity. Antennas and transmitters are presently approaching the optimum. Now it is time to update our receivers to go along with this trend.

My special thanks go to Fairchild Microwave for the use of the test equipment and devices necessary to design this preamplifier. Special thanks go to Will Alexander, WA6RDZ, for all his helpful suggestions. Last but not least, let me thank my wife for typing this manuscript. She says, "It's the most boring thing I've ever typed."

references

1. "The W6FZJ Wide Band Low-Noise Preamplifier," *QST*, (The World Above 50 MHz), November, 1972, page 112.
2. J. Kabell and V.H. Grinich, "Zener Diode Circuits for Stable Transistor Biasing," Fairchild Semiconductor Technical Article TP-8, 1961 (now out of print).
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appendix 1

noise figure

The affect of gain on overall system noise figure can be calculated using the formula:

$$NF_T = NF_1 + \frac{NF_2 - 1}{G_1} + \frac{NF_3 - 1}{G_1 G_2} \quad (1)$$

where: NF_T = overall system noise factor

NF_1 = noise factor of the first stage

NF_2 = noise factor of the second stage

NF_3 = *noise factor of the third stage*

G_1 = power gain of first stage

G_2 = power gain of second stage

Note that all numbers must be in power gain form since the use of dB will result in large errors, especially when low noise figures are used. Noise factor can be converted to noise figure by applying the formula:

$$\text{noise figure} = 10 \log_{10} (\text{noise factor}) \quad (2)$$

Conversely, noise figure (dB), can be converted to noise factor by

$$\text{noise factor} = 10 \text{ antilog}_{10} (\text{noise figure}) \quad (3)$$

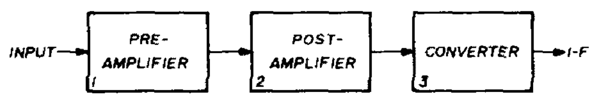
Refer to the illustration below for a typical example. First convert the noise figures (dB) to power factors using **eq. 3**. Then calculate the total noise factor using **eq. 1**

$$NF_T = 1.59 + \frac{4-1}{10} + \frac{10-1}{10 \cdot 16}$$

$$= 1.59 + 0.3 + 0.056 = 1.946$$

$$\text{noise figure} = 10 \log_{10} (1.946)$$

$$= 10(0.29) = 2.9 \text{ dB}$$



dB	NOISE FIGURE, 2 dB GAIN = 10 dB	NOISE FIGURE, 6 dB GAIN = 12 dB	NOISE FIGURE, 10 dB GAIN = 20 dB
EQUIV POWER FORM	NOISE FACTOR, 1.59 GAIN = 10	NOISE FACTOR, 4.0 GAIN = 16	NOISE FACTOR, 10.0 GAIN = 100

Note that the overall system noise figure is 0.9 dB higher than that of the preamplifier alone. Raising the preamplifier gain to 13 dB or dropping the noise figure of the second stage to 4 dB would drop the system noise figure to approximately 2.45 dB – a worthwhile improvement. It quickly becomes apparent that another post amplifier with a lower noise figure may be necessary if you are to approach the low noise figure of the preamplifier. Why pay the price of a low-noise transistor and then lose part of its capability due to a high second-stage noise figure?

appendix 2

zener diode biasing

Refer to **fig. 5**. CR2 is a zener diode and determines the collector-to-base voltage, V_{CB}, of the transistor Q1. The base-to-emitter voltage, V_{BE}, of most low-noise transistors is 0.7 to 0.8 volt and relatively constant. Therefore, the collector-to-emitter voltage, V_{CE}, of Q1 is a fixed voltage which is the sum of the V_{BE} and the voltage of CR2. The current through R3 is given by the equation

$$I = \frac{V_S - V_{CE}}{R3}$$

The current through R3 divides between Q1 and CR2. Ignoring R1, the only current flowing through CR2 is the base current of Q1. Therefore, most of the current flows through the collector of Q1 as I_C. If the dc current gain of Q1 is low, the current through CR2 will increase accordingly. However, if the dc current gain of Q1 is high (as it usually is), the current through CR2 is low and the regulation as a zener is poor. Therefore R1, a 1000-ohm resistor, has been added to the circuit to force at least 0.7 mA through CR2. Since there is usually some base current, a value of 1.0 mA can be assumed for calculations.

As an example, assume you want to operate from a 12-volt power supply with a V_{CE} of 7.0 volts and an I_C of 3.0 mA

$$R3 = \frac{V_S - V_{CE}}{I_C + I_{(CR2)}} = \frac{12 - 7}{0.003 + 0.001} = 1250 \text{ ohms}$$

A 6.2 volt zener would be an appropriate choice for CR2. The collector current can be readily varied by changing either R3 or the supply voltage as indicated in the text. In the circuit of **fig. 4** an additional diode was added in series with the supply voltage for reverse voltage protection. This voltage drop must be subtracted from the power-supply voltage when calculating the value of resistor R3.

solid-state S-meters

These simple solid-state S-meter circuits feature printed circuits that attach directly to the meter terminals

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It's a pleasure to tune in a nice strong S9 +20 dB signal, but it's even nicer if you have an S-meter on your homebrew receiver with which to read signal strengths. The unfortunate thing is that very few S-meter circuits are available, and most of those that are available are carry-overs from the vacuum-tube days. Several popular circuits using either a single device or an IC have the further disadvantage that they are at full scale with no input signal and decrease with increasing signal strength, just opposite to what you would like to see on the meter face.

The most straightforward approach to signal-strength indication is by voltage amplification of the detected audio. Most detected audio signals are in the 10 to 50 mV range, and pushing this up to drive a 5-mA meter movement takes a voltage gain of about 100, plus the added current gain. Neither of these tasks are easy with a single active device.

Commercially available meters with S-unit calibration seem to be limited to two movements, 0 to 1 mA and 0 to 5 mA types. Many modestly priced and military surplus meters also have these same basic movements. In addition, there are a number of specialty meters with removable scales that use a basic 0 to 1 mA movement, and these can be readily adapted for amateur applications.

Fig. 1 and **2** illustrate two types of S-meter PC boards which mount directly to the rear of a meter using the

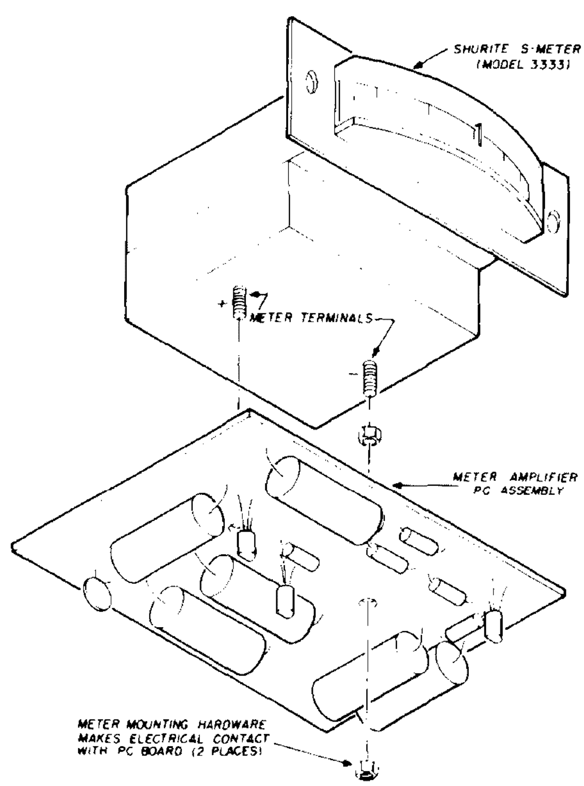

fig. 1. Construction of the 5-mA S-meter circuit shown in fig. 4. Printed-circuit layout is shown in fig. 7.

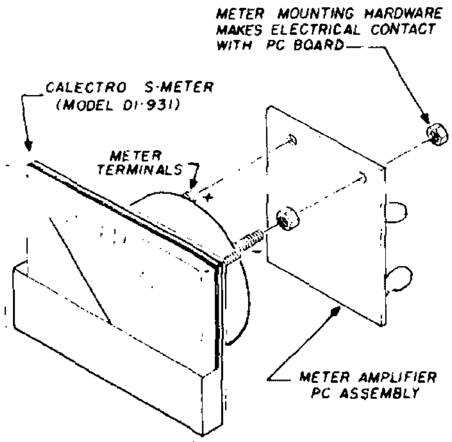

fig. 2. S-meter assembly designed for Calectro D1-931 1-mA meter. Circuit is shown in fig. 3, printed-circuit layout in fig. 5.

meter's plus and minus terminals. This allows free access for panel mounting, separates the meter-amplifier circuits from the rest of the receiver, and provides for easy add-on or modification at a later date.

circuit operation

Fig. 3 shows a schematic for a meter amplifier designed for a 0 to 1 mA S-meter. The fet input provides a high impedance to the detected audio and minimizes loading and distortion problems. The second stage, Q2, is a common-emitter voltage amplifier with a simple positive pulse rectifier for the meter. The 35-μF meter shunt capacitor, C1, filters the rectified audio signal.

Fig. 4 is an S-meter amplifier for a typical 0 to 5 mA meter movement. Similar to the 0 to 1 mA design, the second-stage voltage amplifier, Q2, is followed by an impedance-matching stage, Q3, a simple emitter follower. Performance specifications for both circuits are given in **table 1**.

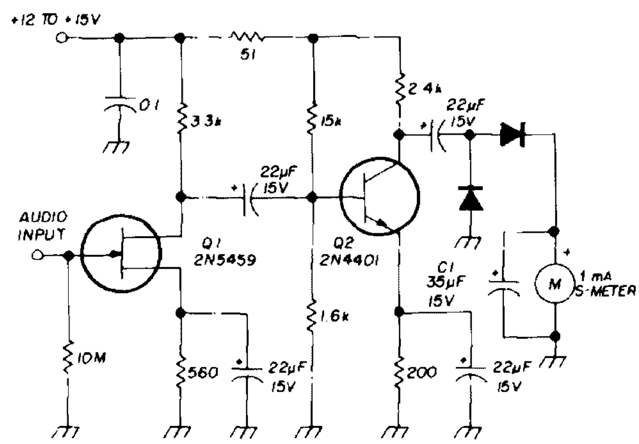

fig. 3. S-meter amplifier designed for a 1-mA meter movement consists of two-stage voltage amplifier and meter rectifier. Printed circuit for this circuit is shown in fig. 5.

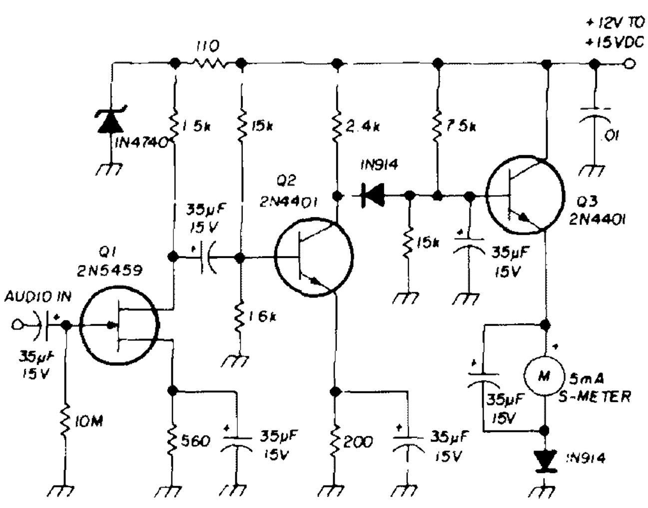

fig. 4. S-meter circuit designed for 5-mA meter movement uses two-stage voltage amplifier with an emitter-follower output. Printed-circuit layout for this circuit is shown in fig. 7.

construction

Fig. 5 and **7** show the PC board layouts and component installation diagrams.* Resistors can be 1/8- or 1/4-watt units although 1/2-watt components can be bent slightly or installed

table 1. Performance specifications of the S-meter circuits shown in figs. 3 and 4.

Audio signal input (S-9)	25-30 mV p-p
Audio signal input (full scale)	50-60 mV p-p
Frequency response	500 Hz to 10 kHz
Input impedance	greater than 1 meg
Power supply	12 to 15 Vdc

*Undrilled printed-circuit boards are available from the author for $1.00 postpaid.

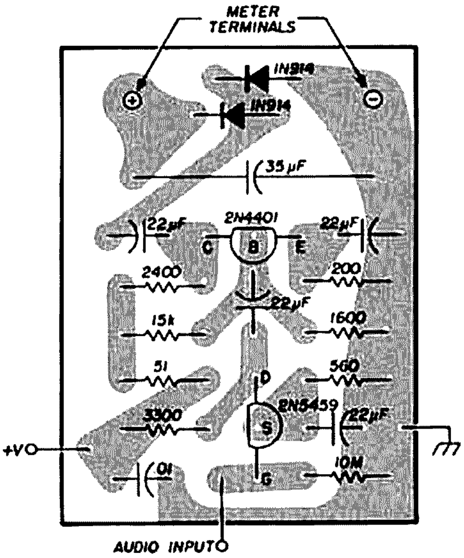

fig. 5. Printed-circuit layout for the 1-mA S-meter circuit. Full-size printed circuit is shown in fig. 6.

vertically on the board to fit. The values for the source and emitter bypass capacitors is arbitrary, and almost any tantalum or electrolytic above 5 or 10 μF will work satisfactorily. An inexpensive dipped tantalum is recommended. The same general capacitance discussion applies to the 5-mA board, but more room is available on this board for the installation of axial-lead components.

An infinite number of transistor substitutions are possible in these two circuits. Almost any audio n-channel, depletion-type fet should work at Q1 by adjustment of the 560-ohm source resistor for maximum voltage gain at the base of Q2. A general-purpose audio npn transistor can be substituted for Q2 and Q3 by adjusting the 15k bias resistor of Q2, and the 7.5k bias resistor of Q3, respectively. In both cases the resistors should be selected for maximum voltage gain to the succeeding

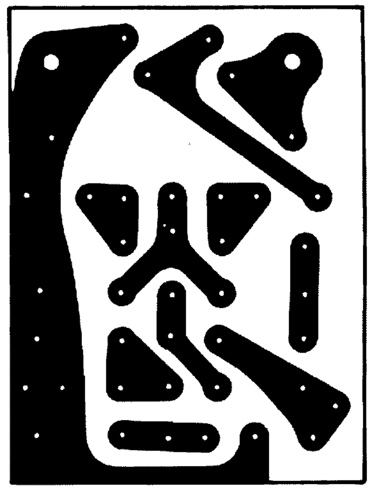

fig. 6. Printed circuit for the 1-mA S-meter. Component layout is shown in fig. 5.

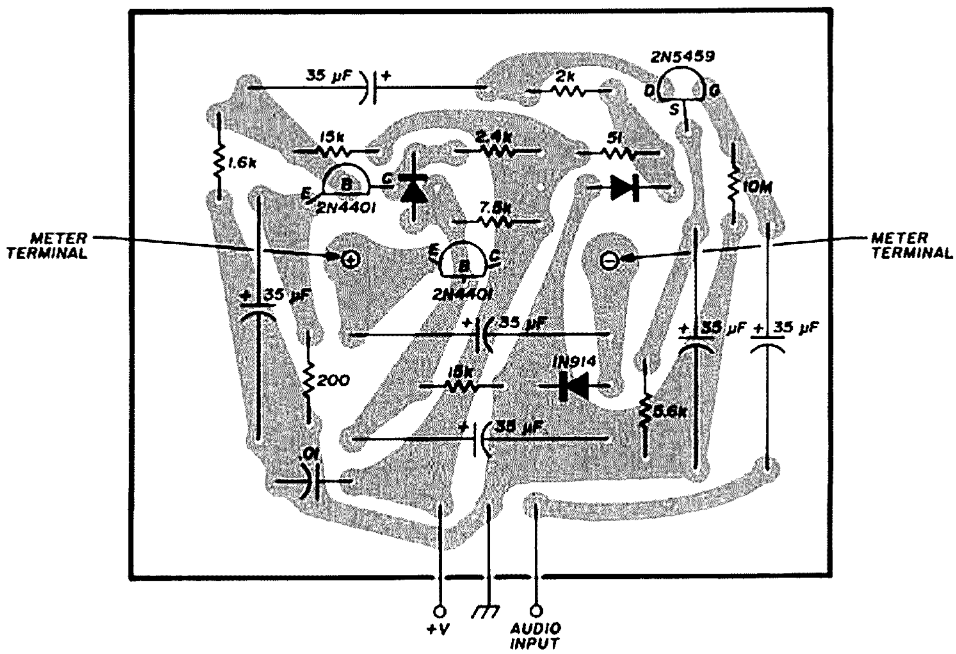

fig. 7. Printed-circuit layout for the 5-mA S-meter circuit. Full-size printed circuit is shown in fig. 8.

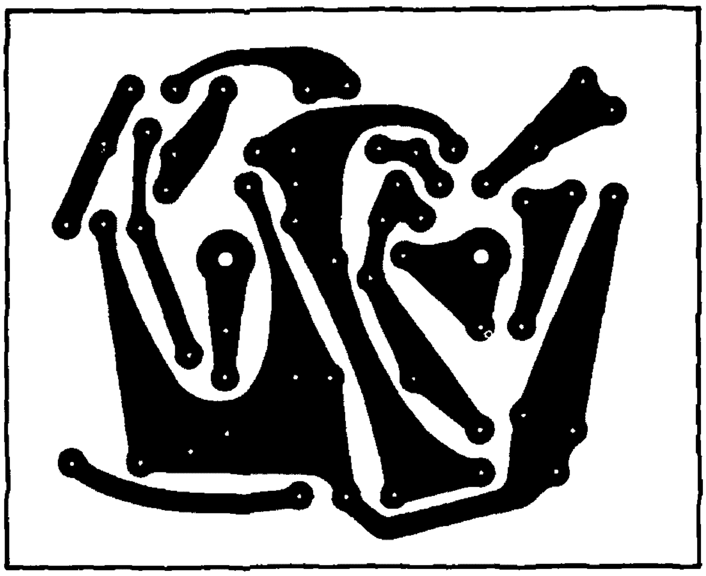

fig. 8. Printed circuit for the 5-mA S-meter. Component layout is shown in fig. 7.

stage or in the case of Q3, selected for maximum meter movement. When selecting bias resistors, it is suggested that a 20- to 50-mV 1000-Hz audio signal be applied to the input of Q1, and that a high-impedance scope or vtvm be used to monitor signal gain.

application notes

Fig. 9 shows a typical application for the S-meter assemblies in a receiver. Since the detected audio level will vary with each receiver, it is necessary to install a sensitivity adjustment. Although this schematic shows a potentiometer, a simple two-resistor voltage divider is adequate. The sensitivity adjustment should normally be set for a meter indication of S9 (approximately half scale) with a modulated 50 μV signal applied to the receiver's antenna terminal. The sensitivity control should be a 50k to 100k unit to minimize loading effects on previous stages.

Since many 0 to 1 mA and 0 to 5 mA meters have different screw terminal locations than those shown here, you may want to use Vector board construction, following the general parts and connection data shown in fig. 5 and 7. An alternative is to locate the PC board externally and use jumper wires to the meter.

If meter motion bothers you, the needle can be damped (or left to free swing) by increasing (or decreasing) the value of the 35-μF meter shunt capacitor.

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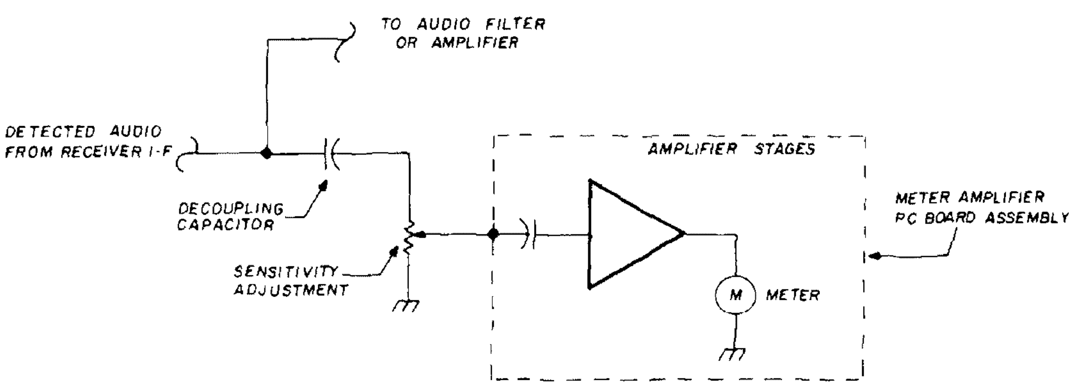
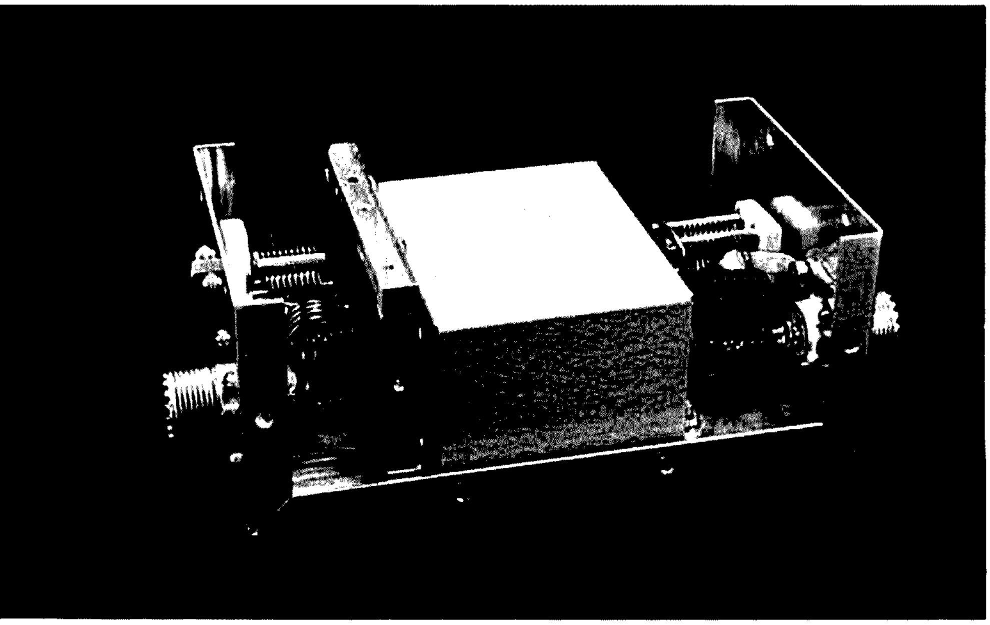


fig. 9. Installing the S-meter amplifier in a typical receiver. Sensitivity pot may be replaced by simple resistive voltage divider, if desired.



high-frequency lowpass filter

A simple, easily duplicated, lowpass filter, with 60-dB attenuation on channel 2 that can be built for under five dollars

Neil Johnson, W2OLU, 74 Pine Tree Lane, Tappan, New York

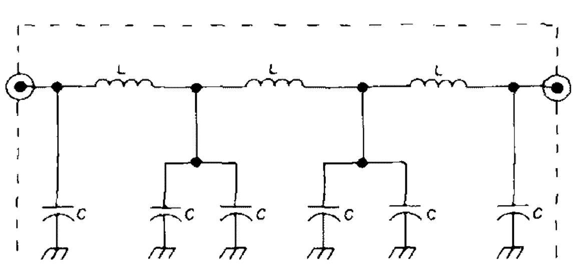
One of the best ways to motivate a casual reader to build a project is to describe something comparatively simple to construct which has more-or-less universal appeal. When such a project is presented along with a solution to any parts procurement problems, the desired result is often achieved: the reader becomes a builder. The following is offered with such a philosophy in mind.

design

Recently a supply of high-quality ceramic capacitors came to my attention at the extremely low price of six for a dollar – they usually sell for three dollars apiece. These are ideal for a TVI lowpass filter being rated at 67 pF at 7500 Vdc and they are small enough to be non-inductive since they're only 3/4 x 3/4 inch (19mm x 19mm). My first thoughts were to use these in a relatively non-critical lowpass filter circuit of the type shown in fig. 1A. This is

merely a combination of three pi-section filters, as shown in **fig. 1B**. No tune-up adjustments would be needed, and the circuit is easily reproduced. That dream went down the drain when the resultant filter showed comparatively low attenuation at TV channels 2 and 3. Unfortunately this is where it is needed most. One could re-design, possibly by lowering the cutoff frequency and thus omitting 28-MHz coverage, but only at the cost of larger, more cumbersome, and more expensive components.

Back to the old reliable. If we elect to keep the good features of the simple filter mentioned above, and then add series-tuned circuits at each end of the filter, we find that most of our needs can be satisfied by the design shown in **fig. 2**. Engineers describe this type of lowpass filter as having M-derived terminating half sections at each end, with two *constant-K midsections*. Most of the construction is non-critical, and when the end sections are tuned to channel 2 (55 MHz) or channel 3 (61 MHz) the theoretical attenuation approaches 60 dB, a ratio of one million to one! Assuming that your transmitter is properly shielded and filtered, this should cure any TVI problems except the toughest ones. As an unintended bonus, this should prove to be a very rugged filter, difficult to burn out even when operated on lines with high standing-wave ratios.



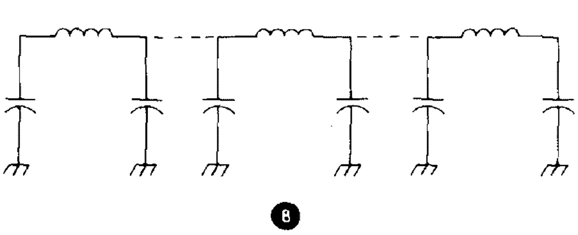


fig. 1. Initial filter design in (A) provided poor attenuation at crucial lower television channel frequencies. Electrical equivalent of the circuit is shown in (B).

chassis

I used an unpainted LMB type 780 "tight-fit" chassis as the basis for this filter. It is made with a great deal of attention, and the close mechanical tolerances employed by the manufacturer should endear this type of chassis to all filter builders. I also tried to simplify the internal shielding problem by using a smaller LMB type 770 "tight-fit" chassis to enclose the center section of the filter. While this is a good design, I felt that even better shielding and harmonic attenuation were called for, at least in the prototype circuit. The additional vertical shield is the result of these considerations.

Personally, I doubt whether such extreme "weatherproofing" is required in all installations. You might simplify your construction initially by omitting the vertical interstage shield, adding it later if needed. In a strong TV signal area I don't believe this shield will be necessary. The ¼-inch (6.5mm) square rods which I attached to the main chassis cover as a precautionary measure are similarly optional.

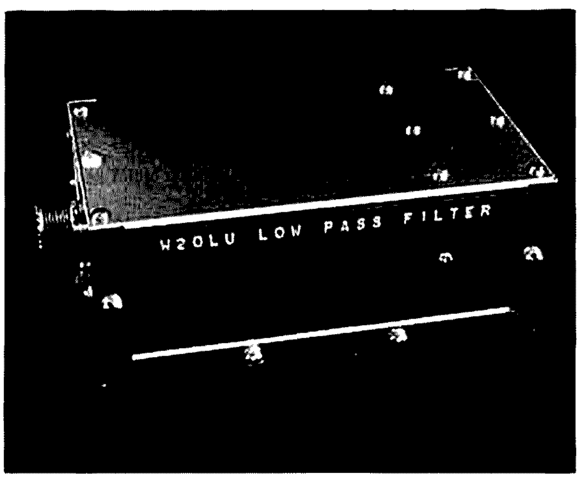
coil winding instructions

All five coils are wound with either number-16 or -14 enameled wire,* using a ½-inch (12.5mm) diameter form as a mandrel. Wind the coils in closewound fashion and leave 1-inch (25mm) pigtails at each end. Bend these at approximately right angles, carefully remove

*Do not attempt to use bus-bar wire or hard drawn copper wire for the coils. These types are prone to spring open to a greater diameter when removed from the ½-inch (13mm) form.

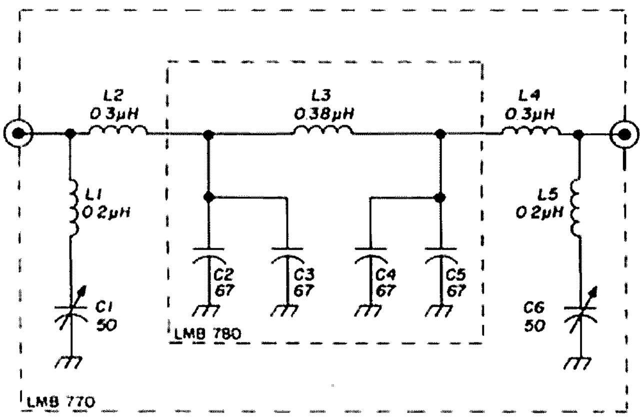
the enamel, and tin the ends. Space the coils uniformly by using a 2-inch (51mm) length of number-14 wire as a gauge, passing it between the turns from one end of the coil to the other. After this is done, make certain that the shape and spacing of the coils is not altered by handling; if this happens, re-space the turns with a piece of number-14 wire as outlined above.

I used number-16 enameled wire for the end-section coils (L1 and L5), since the current carrying capacity of number-16 wire (22 amperes continuous in open air) is more than adequate. I chose number-14 wire for the inductors carrying the throughput (L2, L3 and L4) since I wished to keep losses down. The filter should be capable of handling the output of all but the highest powered amateur rigs, although it was designed to serve the 90% of amateurs who own high-frequency rigs in the 200 to 300 watt input class. All capacitors were checked at 1000 volts ac before installation, and the completed filter was tested at 700 volts ac. A 1-kW input transmitter will have an rf output of 800 watts at most; if this is fed into a well matched 50-ohm coaxial line, the maximum rf voltage will be 200 volts at a line current of 4 amperes.* The filter could probably handle this with no trouble.



W2OLU's lowpass filter is housed in LMB-780 chassis which provides a compact, nearly rf-tight enclosure.

It is not necessary that the mechanical layout be followed, but if you do, success will be easier to come by. One item of note: In the mechanical design of this filter I took a close look at the problem of interstage leakage. When you consider that the coils are only



C1,C6	50 pF APC or MAPC variable
C2,C3 C4,C5	67 pF ±5%, 7500 working volts dc (Centralab type 850S ceramic capacitor, 6 for $1.00 from John Meshna, P.O. Box 62, E. Lynn, Massachusetts 01904)
L1,L5	0.2 μH, 3 turns no. 16 or no. 14 enamelled, ½ inch (13mm) ID, spaced 1/8 inch (3mm) per turn
L2,L4	0.3 μH, 5½ turns no. 14 enamelled, ½ inch (13mm) ID, spaced 1/8 inch (3mm) per turn
L3	0.38 μH, 7 turns no. 14 enamelled, ½ inch (13mm) ID, spaced 1/8 inch (3mm) per turn

fig. 2. Final design: lowpass filter with 42.5-MHz cutoff frequency and theoretical attenuation of TV channels 2 and 3 greater than 60 dB. Filter is enclosed in LMB 780 chassis (5.25x3.0x2.13 inches [134x76x54mm]; components C2, C3, C4, C5 and L3 are enclosed in an LMB 770 box (2.75x2.13x1.63 inches [70x54x41mm]). Construction details are shown in the photographs.

½-inch (13mm) in diameter, the large holes often found in interstage shielding assume noticeable proportions. To keep inductors L2, L3 and L4 from "seeing" each other, I reduced the diameter of the pass-through holes to 3/16 inch

*Standing waves on the line will cause the maximum rf voltage to increase. Editor

(5mm). Also, center inductor L3 was offset so that it was mounted as far as possible from the other two parallel mounted coils. Attention to these seemingly small details makes the difference between a first-class filter and one having only "so-so" characteristics. It is all too easy to nullify part of the potential attenuation of such a filter unless you remember that our goal is to cut down on TVI harmonics by a ratio of a million-to-one!

tune up

When the end sections are first installed, short them to ground with a short strap and dip them to 55 MHz (61 MHz for channel 3) with a grid-dip oscillator. If you lack this simple instrument, I found that if the filter is constructed closely following the original, the two end capacitors are each about 75% fully meshed and at an angle of 41°. This was measured both with a machinist's protractor and the 25¢ plastic type that can be found in stationery stores. If the filter is built as described, the intermediate sections should come out close to the optimum figures (see fig. 3) since the mid-section capacitors are rated at 5% accuracy.

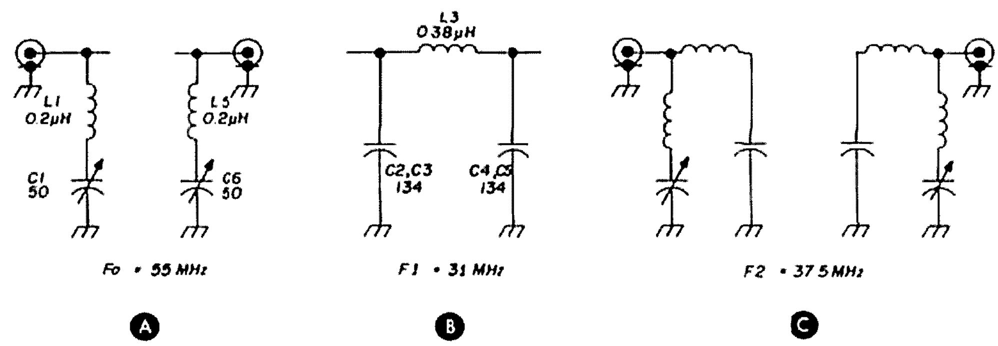
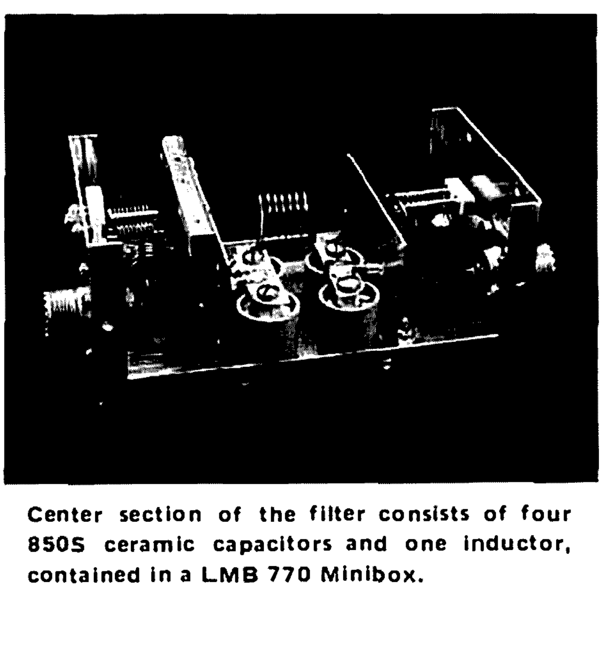


fig. 3. Tuneup data for the lowpass filter. The input and output sections in (A) are first shorted to ground and tuned to 55-MHz (Channel 2 video carrier) or to 61 MHz (Channel 3 video carrier). Center section in (B) should resonate at 31 MHz, and terminating half-sections (C) should resonate at 37.5 MHz. Filter cutoff frequency is 42.5 MHz.



Center section of the filter consists of four 850S ceramic capacitors and one inductor, contained in a LMB 770 Minibox.

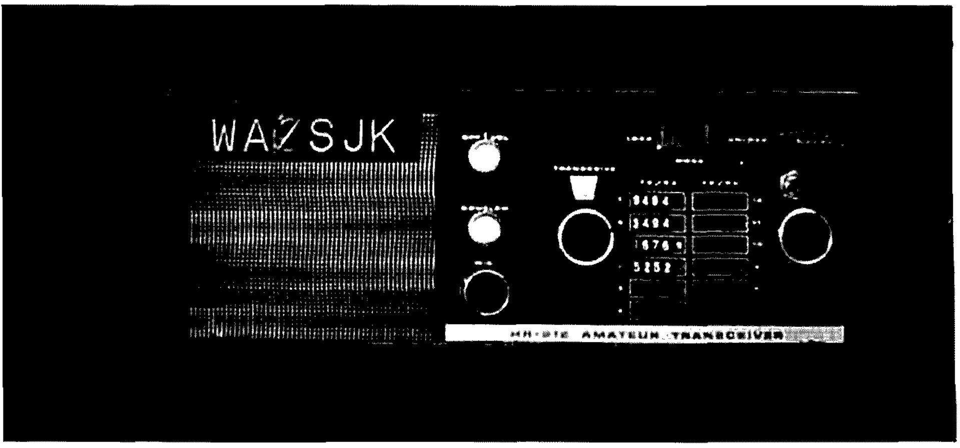
conclusion

Full instructions for dipping and tuning lowpass filters can be found in both the *ARRL Handbook*[1] and the *Radio Handbook*[2]. Follow the drawings closely and the odds are that the TVI beasties will no longer flow out through your coaxial line. Total cost? A pleasant surprise in these days of high inflation. How does five dollars strike you . . . and with spare parts left over!

references

1. *Radio Amateur's Handbook*, 47th edition, ARRL, Newington, Connecticut, 1970, page 585.
2. *The Radio Handbook*, 17th edition, Editors and Engineers, Ltd., New Augusta, Indiana, 1967, page 380.

ham radio



channel scanner

for the Regency HR-212

Construction data for a 12-channel frequency scanner to update this popular two-meter transceiver

Ray Johnson, WAØSJK, Marion, Iowa 52302

If you've been thinking about trading your Regency HR-212 two-meter fm transceiver for one of the scanning models on the market, but find you have more time than money, then this modification may be the solution to your problem. For about $20 you can provide your HR-212 with scanning capability.

Explicit details on parts layout are not included because the circuit was built on a board designed for experimentation and isn't the best possible layout. Although I used wire-wrap for IC interconnections, there's no reason why someone with more time and talent couldn't use printed-circuit techniques. As the photo shows, the mods made to the transceiver front panel don't detract from its appearance.

Since the HR-212 uses diode switching of receive crystals, half of the scanning circuit is already contained in the unit. The circuit described here is essentially an electronic switch which replaces the receive mode rotary switch, S2, on the HR-212.

The scanning circuit operates as follows:

1. The circuit scans 2, 3, 4, 5, 6, 7, 8, 10 or 12 channels, depending on how programmed. These numbers can be changed easily when crystals are added or removed.

2. Scanning will stop and remain stopped on the channel being monitored if:

   A. There is an incoming signal on that channel.

   B. The LOCK switch is in the LOCK position.

   C. Transceiver is in the transmit mode.

   D. The SCAN-MANUAL switch is in the MANUAL position.

Scanning will resume three seconds after the locking stimulus has been removed.

3. If scanning has been locked manually, pushing the STEP switch causes a jump to the next channel. If the scanner is stopped on an incoming signal, pushing the STEP switch will cause the scanning operation to resume until another active channel is encountered.

4. The SCAN-MANUAL switch takes the place of the original mode switch, S3, on the HR-212. When placed in the MANUAL mode, the scanning circuit is disabled and receive frequency follows transmit frequency. (Transmit frequency is always determined by the transceive rotary switch.)

Bottom view of Regency HR-212 showing terminal board and voltage-regulator leads protruding through chassis. Small wires go to scanner board switching outputs, others go to LEDs and crystal-switching diodes (S1A).

construction

Receive channel indication is by means of LEDs, one for each channel

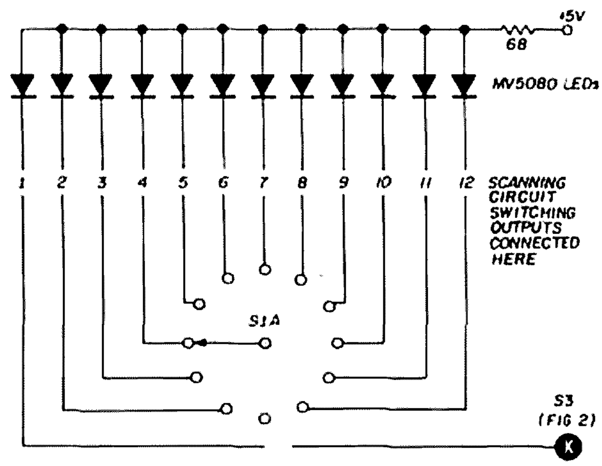


fig. 1. LED board and switching connections. Switch S1A is the transceive 12-position switch on the HR-212. The HR-212 switching diodes are common to this switch (CR501 through CR512).

(fig. 1). The LEDs are placed on a 3x2½-inch (76x64mm) copper-clad circuit board, which is then mounted to the front chassis panel (see installation instructions). These LEDs will then protrude through small holes which are drilled in the front panel of the cabinet. I chose to place the indicators on the right side of the channel identification positions. This method still left enough room for identification using standard ¼-inch (6mm) embossing tape.

The scanning circuit (fig. 2) is constructed on a 2x4½-inch (51x114mm) copper-clad circuit board. Discrete components are interconnected through the copper of the circuit board wherever possible. Wire-wrap sockets are used for the integrated circuits, and connections are made between them and to the

discrete components using wire-wrap terminals which are soldered to the board. Discrete component values can vary to some extent from values given with the exception of Resistor R_T and capacitor C_T. Therefore, don't be afraid

capacitors (fig. 3). This IC comes packaged in a TO-3 transistor case and can be mounted directly to the chassis below the scanner board. Don't neglect to use the bypass capacitors as shown with the voltage regulator as it is impor-

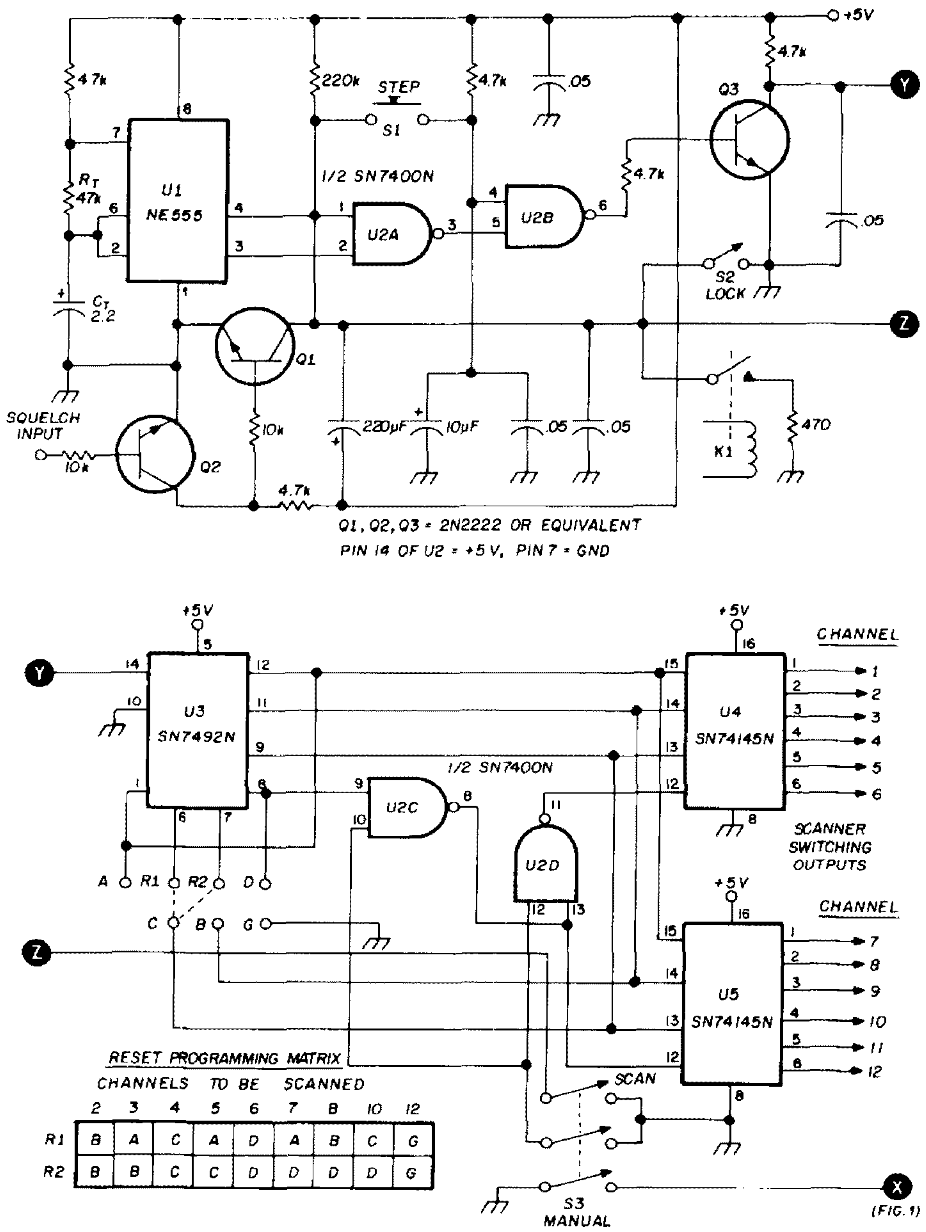


fig. 2. Scanner board schematic. Logic can be changed via the programming matrix to scan any desired number of channels.

to use your junkbox. Almost any npn silicon transistor can be used for Q1, Q2 and Q3.

The 5-volt power supply necessary for the ICs and LEDs consists of an LM335 voltage-regulator IC and several tant to keep rf and transients out of the 5-volt supply.

testing

When the scanner board has been completed, test it with an oscilloscope

or voltmeter before mounting it in the HR-212. Before applying 5 volts, connect the squelch input to a 5-volt point in the circuit and ground pins 6 and 7 of the SN7492 IC (U3). When voltage is applied, a square wave of approximately 6 Hz should be detected on the collector of Q3 (point Y). If no square wave is present, then look for it on pin 3 of the NE555 timer IC (U1). If nothing is there, then check the voltage on pin 4 of U1 and U2; both should read more than 2 volts dc. If the oscillator is working properly, then check the squelch locking circuit by grounding the squelch input. Oscillation should cease abruptly. When 5 volts is reapplied to the squelch input, oscillation should resume in about three seconds.

The scanner board switching outputs are checked as follows: place a 10k resistor between 5 volts and pin 1 of U4 (SN74145N). If the oscillator is running, the voltage on this pin should drop to zero momentarily once every two seconds. Perform this same test for pins 2 through 6 of U4 and 1 through 6 of U5. Now ground pins 10 and 12 of U2 (SN7400N). All switching outputs should remain constantly at 5 volts. If the scanning circuit has passed all the preceding tests, then wash up and prepare for surgery (if you're not shaking too badly).

installation

First remove the dial light from the receive dial (right) side of the HR-212. Then cut the wires off of the receive channel rotary switch (S2 of HR-212) close to the switch. Remove the 12-position rotary switch and replace it with a 3-pole, 2-position rotary switch. Mount the LM335 voltage-regulator IC on top of the chassis (centered in front of the transmitter circuit board). Mount a 12-terminal board or block on the underside of the chassis below the voltage regulator. Connect the wires remaining from the removal of S2 to this terminal board in a sequential manner.

LED board. Position the board in front of the chassis, making sure it clears the transceive dial on the left. Mark and drill holes for the manual-scan switch shaft and the momentary contact step switch (this switch is mounted on the LED board). The step switch will protrude through the hole vacated by the receive dial window on the front cabinet panel. Remove the front panel from the cabinet and drill twelve 3/32-inch (2.5mm) holes for the LEDs. Then line up the LED board and front panel using the hole drilled for the manual-scan switch shaft as a guide.

Clamp the panel and board together and drill the LED board using the front panel as a template. These holes should then be enlarged to 1/8 inch (3mm). Etch a suitable pattern on the board to which the LEDs can be soldered, and make a provision for the 68-ohm current-limiting resistor. When the LEDs are mounted, check to make sure they all work, then secure with epoxy. This is very important, because the MV-5080 LEDs have very delicate leads that will break off if the LEDs have any freedom of movement. Finally, attach thirteen 8-inch (20cm) leads to the board (one for each LED and one for the 5-volt common bus).

Mount the momentary contact switch on the LED board, then carefully position the board on the front chassis panel and mount it on ¼-inch (6mm) spacers. The spacers should be secured to the chassis first, then the LED board should be positioned loosely over both the spacers and nuts. Dress the wires through the front chassis panel and connect them to the proper terminals on the terminal board beneath the chassis. Connect the 5-volt common lead directly to the 5-volt side of the voltage regulator.

Scanner board. Mount four 1-inch (25.5mm) spacers to the chassis and secure them with nuts. Position the scanner board far enough away from the

front chassis panel so that wires can be routed between, then mark and drill mounting holes on the circuit board. Mount the board loosely on the spacers.

Connect twelve wires to the switching outputs using the wire-wrap tool, then connect the other ends to the appropriate terminals below the chassis. The rocker switch on the HR-212 is now used for the locking switch, S2. It

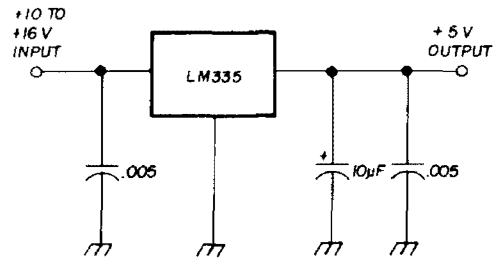


fig. 3. Power supply for the scanner modification uses a LM-335 voltage-regulator IC. Bypass capacitors are important.

should be open in the UNLOCKED position. Two unused, normally open contacts on the HR-212 transmit-receive relay are used for K1. The 470-ohm resistor in series with this relay is optional. Its function is to shorten the delay time of the scanner. This may be desirable if most of your activity is via repeaters.

The squelch input is connected to the HR-212 receiver. Locate R5, a 33k resistor connected directly to the transmit-receive relay; replace this resistor with a 27k resistor. The squelch input line should be connected to the side of this resistor that is *not* common to the relay.

Follow the programming matrix for setting the number of channels you intend to scan. The dashed lines on the schematic show how the scanner would be programmed for four channels. You may wish to employ a switch or switches to facilitate easy programming. Connect the SCAN-MANUAL switch to the circuit board and note that the rotating contact of the TRANSCEIVE rotary switch should also be connected to the SCAN-MANUAL switch at point X. All that remains to be connected now is the power supply. Try the scanner out for awhile before you fasten the board down. The scanning rate can be increased by making C_T or R_T smaller (preferably C_T only).

conclusion

Parts layout for the scanner circuit is not critical. There was enough room on the board for the scanner as well as a tone-burst oscillator, which was designed around a Signetics NE566 function-generator IC. You may wish to lay out the entire project on a printed-circuit board rather than use the wire-wrap method.

The greatest problem encountered with this scanning unit initially was that it tended to go wild during transmit operation due to rf and transients getting into the TTL ICs. This should not be a problem if all bypassing capacitors shown are used. Additional 0.05-μF capacitors can be installed at the V_{CC} input of each IC if this problem persists.

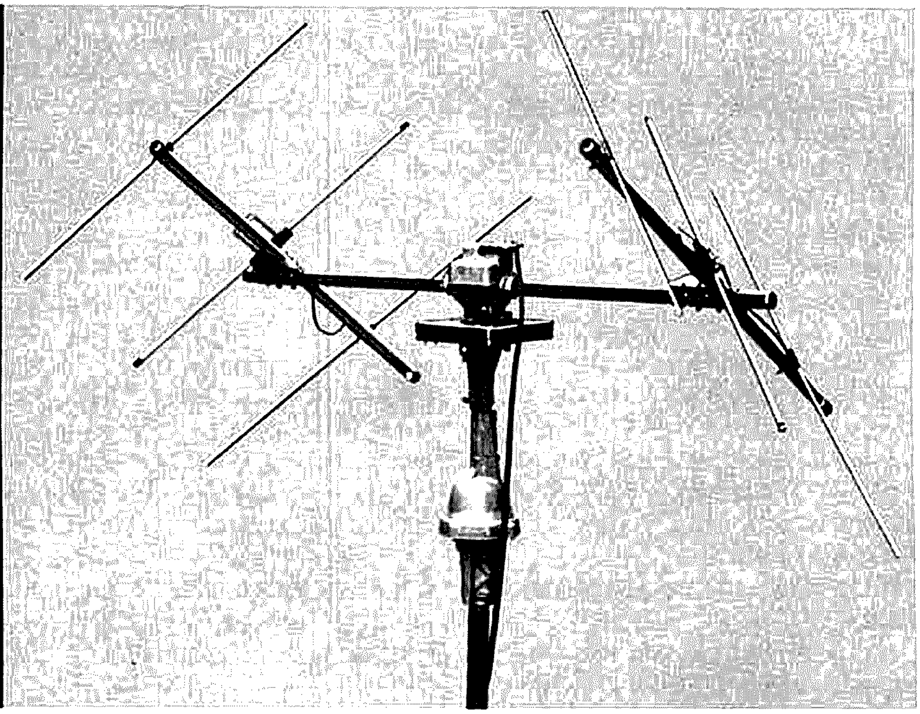
If you don't like the idea of drilling holes in your HR-212 cabinet, you might consider building the scanner as a remote unit. This will require a 15-conductor cable and connectors. The output of the SN7492 counter could be used with a BCD-to-seven-segment decoder to drive a seven-segment display as a means of channel identification.

If you live in an area with a lot of two-meter activity you'll probably find it necessary to improve the selectivity of your HR-212.[1] If you don't, you may find the scanner locked on 146.88 MHz while the received station is transmitting on 146.94 MHz.

reference

1. Paul J. Dobosz, WA8TMP, "Narrowband Modifications for the Regency HR-2 Series of VHF-FM Transceivers," *ham radio*, December, 1973, page 44.

ham radio



low-cost az-el antenna mount for satellite communications

This simple, easy-to-build azimuth-elevation mount will provide added range for satellite contacts

Stuart D. Cowan, W2LX, Box 596, Rye, New York 10580

This article describes a simple, easy-to-build az-el (azimuth and elevation) mount for medium-sized vhf beams, and suggests that experimenting with the angle of elevation of beam antennas and the type of polarization, using the various vhf propagation modes, may yield useful scientific information.

In my az-el antenna mount the azimuth rotor is a TR-44, although a smaller, less expensive unit would work as well with small beams. (Oscar 6 expert K2BZT uses two Alliance U-100 rotors in his az-el mount which handles a 10-element circularly polarized Yagi.) The TR-44 azimuth rotor is mated to the bottom plate of a plywood *sandwich* by an 11-inch (28cm) stub mast and the bottom mast support which comes with CDR rotors. The stub mast can be as long as five or six feet (1.5 to 2 meters) to get the beam antennas higher, but much more than that would

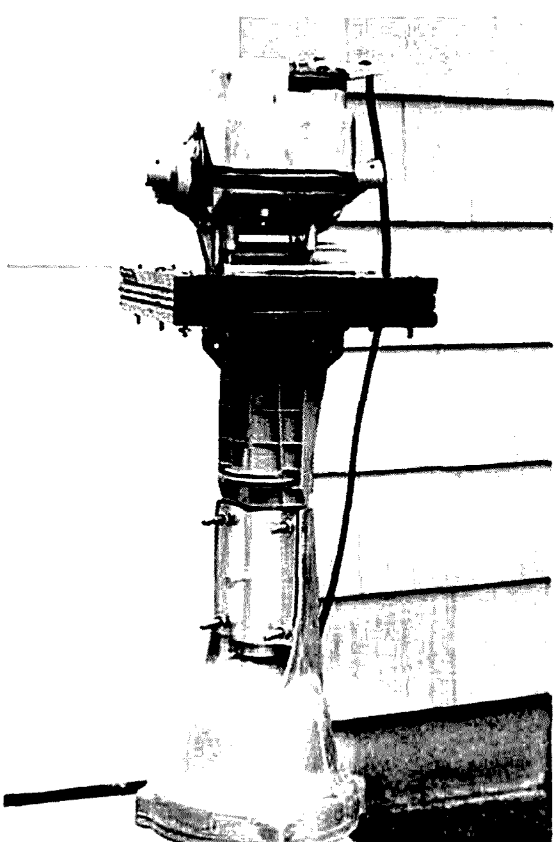
be pushing your luck in ice and wind storms.

plywood sandwich

The plywood sandwich consists of two pieces of half-inch (13mm) plywood 8x9½ inches (20x24cm) and four spacers of the same material which keep the top and bottom plates one-half inch (13mm) apart so that the mounting hardware does not touch. Construction details for the plywood sandwich are shown in fig. 2 and 3.

A unique feature of the sandwich is that the elevation rotor mounts directly over the azimuth rotor which results in balanced downward thrust. Unbalanced weight creates a bending moment which strains both mast and rotor.

fig. 1. Alliance U-100 elevation rotor mounted directly over CDR TR-44 azimuth rotor by means of plywood sandwich and 11-inch (28-cm) stub mast.



Before assembly, the two plywood plates and four spacers should be sanded and given at least two coats of high quality, outside house paint. After the sandwich is bolted together, the two seams are sealed with vinyl tape to keep water out, and the unit is given a final coat of paint. If this sounds too fussy, it is only because I have seen water pour out of waterproof baluns, trickle down the inside of new coaxial cable, split and destroy plywood, and fill a sealed quarter-wave transformer made of hollow tubing. No one will ever know how many weak signals are due to watered-down power!

elevation rotor

The U-100 elevation rotor is attached to the top of the plywood sandwich *using* the mast hardware *supplied* with the unit. Use flat washers and lock washers throughout. Tape the seams of the U-100 to keep water out. The bolts holding together the two halves of the U-100 loosen with time; tighten them every six months.

The U-100 rotor permits the boom holding the vhf antennas to pass entirely through the rotor unit, as the photographs show. Blonder-Tongue Prism-Matic rotors offer the same advantage. This is a must to achieve proper weight balance. If a single crossed Yagi is mounted on one side of the elevation rotor, a counterweight should be mounted on the opposite side.

The cross boom is a 5-foot (1.5-meter) wooden pole from a beach umbrella in order not to detune the Yagis, which should be mounted as closely together as possible. A metal boom could be used, but I prefer a strong, well painted wood boom for proper decoupling.

The two 3-element Yagis are mounted at 90° to each other, and 45° to the boom. One beam could be mounted

vertically and the other horizontally, but that posed a problem in this case because of the beam hairpins and baluns. The stub mast must be long enough to permit the beams to clear the supporting structure when the beam is rotated at full 90° elevation.

elevation control

A word of caution: before installing your masterpiece on top of a tall tower or high roof, conduct a dry run on the ground where you can easily, "correct any malfunction," as they say in the electronics industry. (The first time I elevated my antennas they pointed down into the driveway.)

Calibrating the elevation control is simple. I use east for the horizontal position and north for 90° elevation (straight up). A paper elevation calibration chart is pasted to the control unit between E and N.

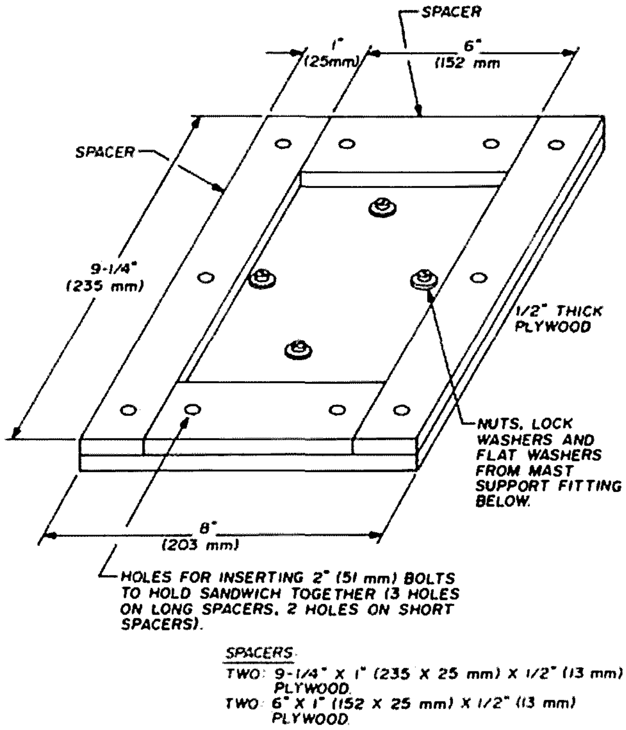


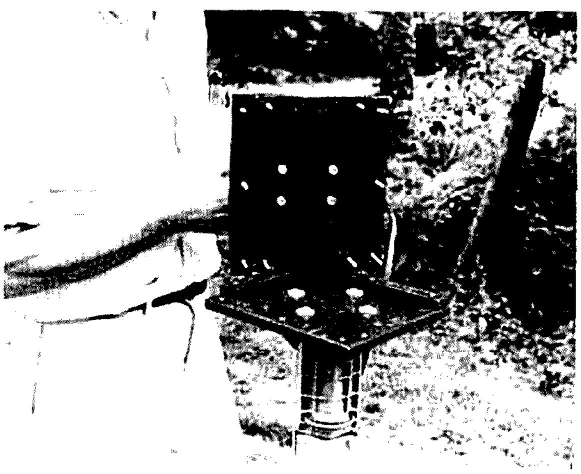
fig. 3. Construction details for the bottom plate of the plywood sandwich. Top plate is similar (see fig. 2).

use with satellites

While you do not need a beam, or an az-el mount, to work successfully through a satellite, using a good beam on such a mount gives superior results with less power. And the tests you can make are very educational. The Oscar satellites give you an opportunity to do something *which amateurs have never before been able to do* – listen on the downlink to your own signal coming back to you from a distant point (outer space) so that you can instantly hear the results of changes in power, types of antennas, and beam headings and elevation angles. There's no guesswork, no baloney, no inflated reports!

With a satellite, those adventuresome souls with a high-gain beam on an az-el mount are the first stations to access the bird on its pass, have the best signals (if they keep the beam heading in the right direction, at the right elevation), and work through the satellite up to the last possible moment before it drops over the horizon. On the other hand, an az-el mount keeps you a lot busier on a satellite pass than a fixed antenna so if you are the nervous type, you probably

fig. 2. Details of sandwich made of ½-inch (13mm) plywood, well painted against weather. Bottom plate bolts to mast support for TR-44 or Ham-M rotor. Top plate bolts to U-100 rotor. Bolts fasten plates together, separated by four plywood spacers. Dimensions are shown in fig. 3.



would be better off to stick to a ground plane. Another possibility is the automatic az-el control system described in the January issue of *ham radio*.[1]

vhf propagation

There are numerous types of vhf propagation to explore and much to learn about over-the-horizon vhf/uhf paths, most of which exhibit very unpredictable behavior. Here is ideal territory for the purposeful amateur seeking a chance to contribute to scientific knowledge.

For example, does the angle of elevation of an efficient multielement beam with a relatively narrow main lobe affect vhf/uhf propagation which results from ducting, sporadic-E, temperature inversion, obstacle gain, aurora reflection, tropospheric scatter, ionospheric scatter, etcetera? If it does affect propagation, then in what way?

It is already known that the angle of elevation is important in satellite and moon reflection communications. Moonbounce expert W6PO reports that once an antenna is high enough to be free of ground effect, elevation control is a key factor in meteor trail reflection, making possible short-range communications (under 500 miles), such as between California and Nevada. Antennas are tilted upward and aimed at a meteor shower over northern Oregon so that the angle of reflection equals the angle of incidence on the reflection path. *Presto* – meteor reflection contacts over short distances!

polarization

Besides experimenting with the angle of elevation, worthwhile tests can be conducted comparing horizontal, vertical and circular polarization (right- and left-hand). W6PO suggests that tilting an array *sideways* – halfway between horizontal and vertical polarization – to achieve "lopsided polarization" might lead to interesting results.

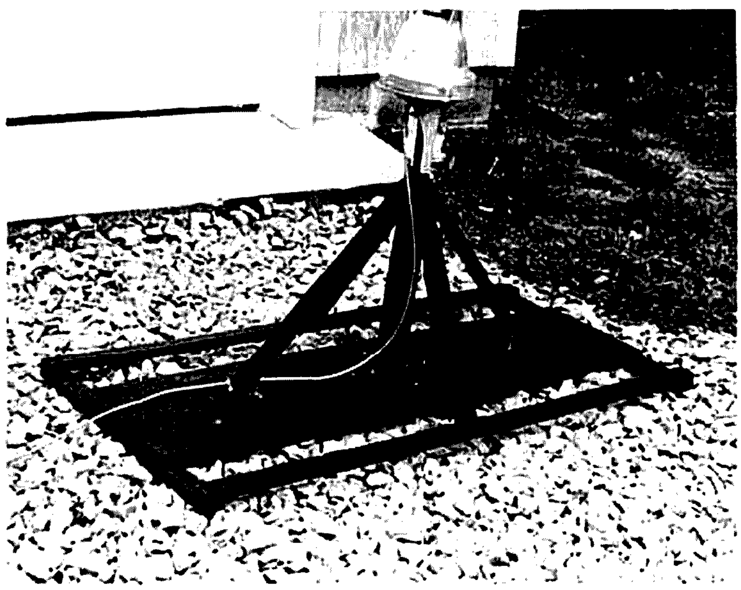


fig. 4. Assembly, supported by four braces, mounted on 2x4-foot (60x120cm) plywood frame for fastening to flat area. TR-44 rotor can be mounted on regular tower. Length of mast from TR-44 to U-100 should not exceed 5 to 6 feet (1.5-2 meters).

Circular polarization may be produced with a helical beam, or with two Yagis, one of which is fed 90° out of phase with the other. The Yagis may be on the same boom, or separate as shown here. With Yagis, the transmission line to one beam is an electrical quarter-wavelength longer than the line feeding the other beam, thus imparting a minute delay in the power fed to the second

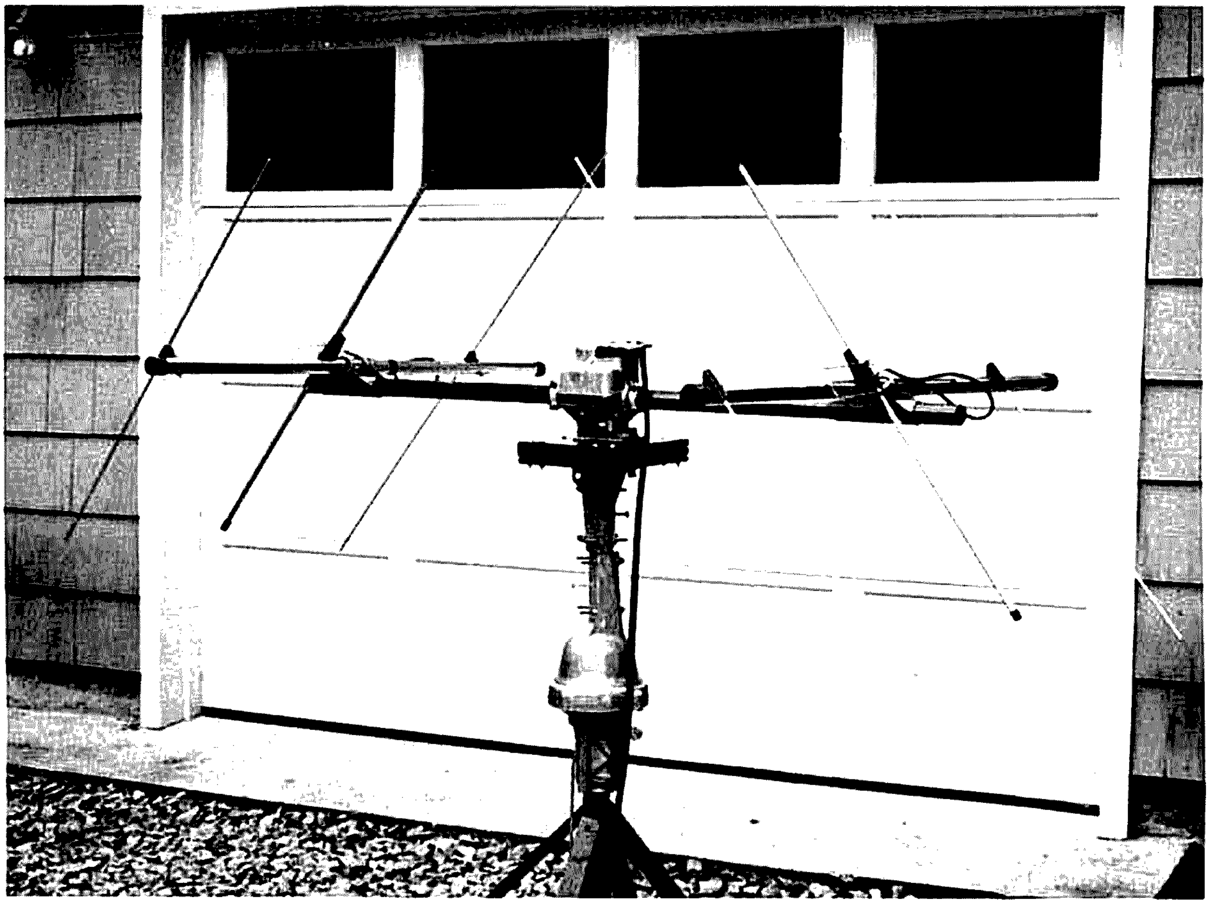


fig. 5. Az-el mount assembled using two 3-element 144 MHz Yagis at 90° to each other for circular polarization. Cross boom extends through U-100 rotor. Feed line and phasing harness not shown.

beam. Data on circular polarization may be found in references 2 and 3.

The results of these tests can be recorded by a reliable observer at the other end of the path, preferably using a tape recorder, or better, by having that amateur patch your signal into the telephone line so that you can actually hear the effects of the changes you make.

summary

Vhf and uhf territory, 50 MHz and higher, is one of the few remaining frontiers where pioneering amateurs still have an opportunity to contribute scientific discoveries in the honored tradition of the amateur service (the discoveries have been a bit sparse of late).

Except for satellite and moonbounce communications, little is known about the effects of changes in angle of elevation of antennas, and their polarization, on the propagation and reception of vhf and uhf signals. Who can tell what awaits the determined amateur with a pioneering spirit?

Thanks are due K2BZT, W6PO and W6SAI for data in this article, and to WA2ECC for the photographs.

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2. Katashi Nose, KH6IJ, "Crossed Yagi Antennas for Circular Polarization," *QST*, January, 1973, page 21.
3. Herbert S. Brier, W9EGQ, William I. Orr, W6SAI, *Vhf Handbook for Radio Amateurs*, Radio Publications, Inc., Wilton, Connecticut 06897, pages 224, 287 *et seq*.

ham radio

programmable calculators

for solving engineering problems

An example of the versatility of these machines plus some expert advice on their use

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During the last few years the use of small electronic calculators has become increasingly popular in broadcast and communications engineering. The small units available, even the $19.95 housewives' grocery-store special, offer a speed of operation and portability far above that of the slide rule. Almost overnight the field of mechanical calculators has been wiped out by progress.

In keeping with progress, a new breed of calculators of the minicomputer type has appeared on the market and several machines of reliable manufacture are now available. They are within reach of the average budget and will prove themselves in the first engineering problem calling for a repetition of calculations to be made in some orderly sequence. These machines are known as programmable calculators: they can accept one or more programs and can process them from a series of memory registers. Numbers may be recalled, changed, or restored at will in these registers.

programmable computers

The word "programmable" is apt to conjure up some scary ideas in the mind of the casual reader. Let's see what the word means. We may start out and store in the memory registers those quantities that are known or which we'll want to process, doing all this before we load the program into the machine. If desired, a sample calculation may be made at this point to acquaint ourselves with the routine, but this is not really necessary. Better that we go into LOAD, then proceed as if we were making an ordinary calculation, manipulating the keys to work toward an answer. Then we go back into the RUN mode. The program is now stored and we may insert new values into the memory registers and give the machine the go-ahead to process the new data.

The beauty of such an operation is that it doesn't have to be entered in a precise machine language such as BASIC, FORTRAN IV, or COBOL. In these languages, the big sophisticated machines must be commanded in an exact routine, using precise statements, and using exact names for what is to be

done. In the operation described here, the user simply goes through a series of key manipulations as if he were operating an ordinary electronic calculator.

A relatively new machine, the Compucorp 324G,* will perform such routines and run a very high program-step-to-memory ratio, presently 8:1, and will present 10-digit results! The sophistication of the programming is left to the imagination of the user. The machine actually processes to the 13th digit although only presenting 10 – with no apologies being offered to users of the IBM-360 or to equally serious programming with ordinary single-precision results.

elementary programming

Without attempting to go into an elaborate computer course, one or two no-no functions apply equally to the Compucorp 324G and the big IBM jobs; and the first of these is multiplication or division by zero. This is another way of saying that the memory registers should be loaded *before* inputing a program, else the idiot symbol, such as *E----*, will appear. There are ways of getting around this and the procedure will be explained later. Another no-no is attempting to take the square root of a negative number, in which case the result is identical. In both cases the RESET button is convenient, and forgiveness is instant.

Before explaining how you might wind up in either of the above situations, a brief explanation on how to form a loop is given. We have always been accustomed to using the equal sign (=) in expressions such as 2 + 3 = 5. Think of this idea in another context: "I have taken 2 and added 3 and replaced it by 5." Both statements are the same since I am no longer interested in the 2 or the 3, but in my future thoughts I now have 5. I could do the same thing on the machine if I had a 2 in the program, recalled a 3 from a memory register, took the sum, and placed the 5 back in the memory register formerly occupied by the 3. The next time around I might add 2 again, and I would recall the 5 from the register and put the sum 7 back into the same register, and so on.

some examples

The above was a simple exercise in repetition, which may be included in a program as a stepping function. A typical example would be in the design of a directional antenna where you want to obtain a value of field intensity for every five degrees of azimuth. The above routine would be placed either at the beginning or the end of the program, storing the azimuth value in one of the registers and including the quantity for each successive step either in the program itself, or in one of the other registers. Each time the START button is pressed, the value in the register containing the azimuth would step forward, and the calculator would compute

*Marketed in the United States and Canada by Monroe Business Machines

Several years ago *ham radio* published a rather whimsical article about how computers might someday influence amateur radio.* The author made a prediction that has now become a *reality: low-cost calculators for home use.* In a later issue we presented an article by another author, who described a method of circuit analysis using a canned program, ECAP, but which required large machines such as the IBM 1620.† Here's yet another offering on solving electronic design problems using the computer. Only this time the machine is one of the relatively low-cost minicomputers known as a programmable calculator. Author Aylor leads you through the action in a step-by-step fashion (with advice on avoiding pitfalls) from initial programming to the final result – all on the new Compucorp 324G. Time has indeed marched on. Editor

*Louis E. Frenzel, W5TOM, "Computers and Ham Radio," *ham radio*, March, 1969.

†M.A. Ellis, K1ORV, "Computer-Aided Circuit Analysis," *ham radio*, August, 1970.

the new value of field intensity at the new azimuth. Don't forget that you can recall the stored number from your controlling (azimuth) register as many times as you wish with the RECALL button, and the value will not be destroyed until you use the STORE key.

The stepping function, which is incorporated in a loop, is by no means confined to additive quantities. In the directional antenna example you could have as easily started at 360 degrees (or 180 degrees for that matter) and used the subtractive form of stepping; i.e., taking away 5 degrees each time the function was repeated.

Stepping to form a loop is by no means confined to any specific part of an equation. For convenience, the Compucorp 324G micro-computer has a set of parenthesis or brackets and you can form a stepping loop within these boundaries if desired. There is one additional no-no associated with the use of these: you must always use the same number of closing parentheses as used for entrance. Failure to do this will get you that great big error *E*---- signal, and only use of the reset button will clear the machine. Once that signal comes up, the reset button is the only key on the board that will work. Moral: get it in right next time.

Compucorp®

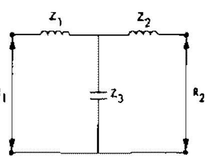
Page 1 of 2

PROGRAM: DERIVATION OF BRANCH REACTANCES FOR A MATCHING T NETWORK (SEE ALSO 8807471-B)

Program No. 8807471-A
Models 324
Date 3/74
Author Aylor

DESCRIPTION

This program derives branch reactances in a T network, to match given input and load resistances for a given phase shift. Thus, one may change ratios and phase shifts as desired to obtain network characteristics.



$$Z_1 = -j\frac{R_1 \cos\phi - \sqrt{R_1 R_2}}{\sin\phi}$$

$$Z_2 = -j\frac{R_2 \cos\phi - \sqrt{R_1 R_2}}{\sin\phi}$$

$$Z_3 = -j\frac{\sqrt{R_1 R_2}}{\sin\phi}$$

where:

R_1 = Generator resistance, ohms
R_2 = Load resistance, ohms
ϕ = Phase shift, degrees
Z_1, Z_2 and Z_3 are leg reactances in ohms

EXAMPLE AND TEST PROBLEM:

Find the values of Z_1, Z_2, and Z_3 required to match a generator (R_1) whose output impedance is 50 ohms to a load (R_2) of 75 ohms for the resistive case where $\phi = 70°$.

1. Follow loading instructions on reverse side
2. Enter / Press

Enter	Press	
50	ST 1	Input
75	ST 2	Output
70	ST 3	Phase shift
	S/S	

3. -65.16 displays

Press	Read	Description
RCL 7	46.96	Reactance of Z_1
RCL 8	37.86	Reactance of Z_2
RCL 9	-65.16	Reactance of Z_3

Program Number 8807471-A Page 2 of 2

OPERATING INSTRUCTIONS

Loading the Program

1. Enter 1 ST 3
2. Switch to LOAD, PROG I
3. Key in program
4. Switch to RUN
5. Set DP as desired

Running the Program

6. Enter and store the variables as required:

Enter	Press
R_1 (ohms)	ST 1
R_2 (ohms)	ST 2
ϕ (degrees)	ST 3

7. Press S/S Z_3 displays
8. To Read / Press

To Read	Press
Z_1	RCL 7
Z_2	RCL 8
Z_3	RCL 9

Note: This program yields the reactive values in ohms. Program B determines actual values in microhenries (H) and picofarads (pFd) to construct the network, retrieving values in registers 7, 8 and 9.

REGISTER

0	1	2	3	4	5	6	7	8	9	F2
	R_1	R_2	ϕ				Z_1	Z_2	Z_3	

PROGRAM STEPS

1	RSET	41	RCL
2	RCL	42	6
3	1	43	=
4	X	44	DIV
5	RCL	45	RCL
6	2	46	4
7	=	47	CHSN
8	SQRT	48	=
9	ST	49	ST
10	6	50	8
11	RCL	51	RCL
12	3	52	6
13	SIN	53	DIV
14	ST	54	RCL
15	4	55	4
16	F2	56	CHSN
17	ST	57	=
18	5	58	ST
19	X	59	9
20	RCL	60	S/S
21	1	61	
22	=	62	
23	-	63	
24	RCL	64	
25	6	65	
26	=	66	
27	DIV	67	
28	RCL	68	
29	4	69	
30	=	70	
31	CHSN	71	
32	ST	72	
33	7	73	
34	RCL	74	
35	2	75	
36	X	76	
37	RCL	77	
38	5	78	
39	=	79	
40	-	80	

fig. 1. Part A of a program for use with Compucorp 320-series programmable calculators to solve a typical problem: finding component values for a T-network. Page 1, left, presents the problem, shows data to be entered into the machine, and shows the network reactances obtained from the program steps on page 2, right*.

*The equations for Z1, reactance of the input leg; Z2, reactance of the output leg; and Z3, reactance of the shunt coupling mesh; are as presented in F.E. Terman's *Radio Engineer's Handbook*, McGraw-Hill, New York, 1943, page 212 (equation series 113).

using parenthesis

Here's a specific example of a short program with an internal loop. In this instance, the stepping function will be

enclosed within parenthesis, which stipulate the power to which a number will be raised. This is the old "double-each-day" routine. You've heard the one of going to work for only a penny a day and having your wages doubled every day and so on until you make it rich. The equation for this exercise is day's pay = $2^{(N+1)}$ x 0.01. Let's see how to form a loop with the Compucorp 324G.

First go to RESET, which clears the display, then switch to LOAD and the machine is ready to receive the program. Then proceed as follows:

action	explanation
Punch 2	Installs the integer to be raised to (N + 1) (program starts).
A^x	Instructs the machine that the next expression will be raised to a power.
(	Entrance parenthesis.
1	Incrementing step, once per day.
+	Addition command.
RCL_n	Prepares machine to go into memory register on next stroke.
0	Memory register stored under key 0, reads, and presents sum.
=	Prepares for storage.
ST_n	Alerts for next stroke, which will put selected number under next key punched.
0	Destroys that which was previously in register 0 and puts in a new quantity, N + 1, which is the stepping function.
)	Closing parenthesis. The quantity (N + 1) is now poised and ready for action because you conditioned the machine with A^x earlier.
=	This is the go-ahead for the machine to raise 2 to the (N + 1) power. Remember, the original was raised to an increasing power each day.
X	Multiply command. You must convert whole numbers to pennies and dollars.
.01	One hundred cents per dollar.
=	Presents final result.
START/ STOP	Halts program.
RESET	Clears display to all zeros.
RUN	Terminates and conditions computer for RUN mode.

Now let's see what we've done. The calculator is capable of storing up to 80 steps or operations, and the display presents a running account of each time you press the key. Although there are only 17 *apparent* actions before the switch to RUN, the above storage actually consumed 20 of the 80 available steps. Why? Because, near the end of the program, entering .01 to convert cents to dollars required three key operations – the decimal, the zero and the one. The last step is actually a switch operation and does not count as a step.

We are now ready for a step test. For exactness we store a zero to obliterate anything remaining from the program entry operation, then press START/STOP and the machine momentarily displays a 1, meaning the first day has passed. The machine pauses and presents .02. Press START/STOP again and the machine displays a 2, pauses, then presents .04, and so on. After pressing START/STOP twenty-four times, the answer will be 167,772.16, or in hard Uncle-Sam-type greenbacks, $167,772.16 – not bad for a day's pay after only twenty-four days on the job!

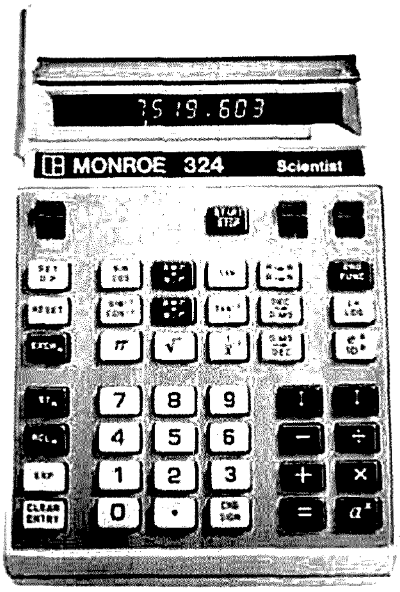
additional looping forms

The example above was a rather simple one in a relatively common exercise. Earlier we used an example of destructive replacement in the 2 + 3 = 5 procedure, and that is exactly what we are doing here. In this example we installed the stepping increment (the fourth action), and on the sixth and seventh action we retrieved the quantity stored in register 0. The moment we punched the EQUAL button, the program added 1 to the quantity retrieved. Thus added, we are no longer interested in the quantity stored in register 0 so we use the storage key (ST_n) and store the new, added value in register 0.

To review, you went into register 0, retrieved what was there, added 1 to it, used it in the calculation, then destroy-

ed it by placing a new number as a replacement into that register, which previously contained the old number. All you did was replace the old value of N with a new value, defined as N + 1.

The rest is simple: put in a closing bracket and you have the portion of the equation shown as a power all set to go.



The Monroe 324G Scientist programmable calculator used by the author.

The calculator has been patiently waiting, poised since you told it that you were going to raise 2 to a power in the first and second steps of the program. Obviously, all that is necessary now is to punch the equal button – calculation gets under way immediately. The rest of the procedure places the decimal point, presents the result, and stops the calculation. Omission of START/STOP, a common mistake, will allow the machine to run wild until it hangs up on some value with more than 99 values or figures presented as an exponent in scientific notation.

There are other variations that are somewhat more sophisticated but easily worked on the machine. Suppose you want to construct a map and keep the longitude constant while walking up the curvature of the earth with a stepping value and reach a distance and bearing for each intercept – or suppose you want to construct a tower guy anchor (actually a large block of concrete) and given the weight requirement, you must know the most economical dimensions for placement. Or suppose you must design a transmission line-to-antenna network while solving for all phase shift values. The rule is the same: store it, recall it, and sum it in presentation. Store the presentation in the same register that kills the old value by destructive replacement and let the remainder of the program do the spade work, then go on from there.

There are a few other routines that will become apparent as the user gets more familiar with the use of the machine. For example, the Wayne-Kerr balanced impedance bridge, which is manufactured in Great Britain, gives a reading of the equivalent parallel circuit in terms of millimhos conductance shunted by equivalent picofarads susceptance, and the instruction book gives the equations for conversion. The equations are basic, requiring the conversion of the susceptance to reactance at a specified frequency.

Now, the condition where the susceptance dial reads zero picofarads is very real and means that the load is purely resistive, a function of the conductance. If, during the course of the factoring, the calculator encounters a zero in the denominator, it will latch up on the error signal. The way to avoid this is to input a small additive constant, such as 0.1×10^{-30}, into the capacitance value within the program. The machine adds the zero capacitance to the 0.1×10^{-30} picofarads residual, and the sin committed never shows on 10-place results. Of course, it would

have been easier to have inverted the conductance with the 1/X button and ignored the capacitance reading. But the objective is to achieve a program that will work with any list of data you use, with an answer (meaning no error latch-up) on anything you insert, and this means *no* exceptions.

Although taking the square root of a negative number was introduced as a no-no at the beginning of this article, there may be an exception in certain problems where you are stepping one or more variables in a multiplication form and wish to compare the product with a constant that is the criteria beyond which you cannot go. This constant is stored in another register. You then arrange to subtract the product from the constant and take the difference, then promptly take the square root. When the calculator attempts to go past the boundary condition – you guessed it: *error*! It's easy to reset and then recall the values that caused the latch-up. Note that earlier mention was made of dimensioning a large block of concrete to use as a guy anchor for a tower. With concrete costing $150 to $200 a cubic yard in place, overdesign can be costly and underdesign can be disastrous!

a typical program

The discussion that follows is specifically arranged for the Compucorp 324G minicomputer and consists of two programs that can be stored under a single bank of memory registers. The programs are typical of problems encountered in electronics; in this case it's desired to find the inductance and capacitance of a T-network between a generator (usually 50 ohms but subject to the user's choice) and a load such as an antenna, where any impedance may be encountered.

Both programs (**fig. 1** and **2**) are based on standard handbook equations. Program 1 yields the impedance of the three arms of the network. Program 2, into which the desired frequency is entered, gives the network values in μH and pF. In running the programs it's easy to plug in variables, store a value associated with program 1, then give the machine the go-ahead by pressing the START/STOP key. Then you switch to program 2, the machine retrieves the impedance values stored in program 1 and, using the frequency value entered in program 2, converts and displays the network numbers in microhenries and picofarads. Stepping is not included, but plenty of room exists in program 1 at step 60 to recall any values from the first three registers and add the steps in place of S/S (START/STOP) using the same old rule: recall, add (or subtract), store, punch S/S, and be prepared to process again.

analysis

Now let's see what went on in part 1 of the program. The expression under the radical was recalled and multiplied, the square root extracted and put to rest in register 6 at step 10 (**fig. 1B**). Then you recalled the phase shift from register 3, took the sine and promptly stored it into register 4 at step 15 for later use. Next you obtained the cosine with the second function (F2) button and stored it into register 5; but while it was available, you multiplied by the input impedance from register 1 and hit the equal (=) button for the product, then subtracted from the square-root quantity you had stored into register 6 at step 10.

At step 28 you came back for the sine you put into register 4 and divided, switched the sign in step 31, and put the input leg of the network into register 7 on step 33. The same process, with the stored quantities (sine Ø and cosine Ø) and the value under the radical, is repeated, which takes you up to step 50, yielding the output leg impedance. The shunt, or mesh, leg value is obtained by dividing the number under the

radical by the sine, which is entered into register 9. Registers 7, 8 and 9 now have input, output, and shunt leg values respectively; program 1 is concluded with START/STOP; and you switch back to RUN.

Program 2 is for those allergic to reactance tables and only requires 40 steps. The microcomputer is switched to eters in impedances and phase shift. With further recalling, registers 7, 8 and 9 have the reactance values, and recalling 4, 5, and 6 gives you the design value of each component. Don't forget register 0 down in the left-hand corner, which contains the frequency.

At this point, if you want to make changes in any parameter in registers 0,

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Page 1 of 2

PROGRAM

DERIVATION OF ELEMENT VALUES FROM REACTANCE VALUES FOR A T NETWORK (SEE ALSO 8807471-A)

Program No 8807471-B
Models 324
Date 3/74
Author Aylor

DESCRIPTION

This program derives the component values in a T network from the given reactance values independently, or as calculated by Program A.

FEATURES:

- The program is capable of specifying component values for any frequency desired (frequency entered in MHz).
- All unit conversions to microhenries and picofarads are provided for in the program.

EXAMPLE AND TEST PROBLEM:

Having determined reactive values for transfer from 50 to 75 ohms at ϕ - 70° to be Z_1 = 46.96, Z_2 = 37.86, and Z_3 = -65.16Ω, find network component values for f_o - 1.25 MHz.

1. | Enter | Press | Description |
|---|---|---|
| 1.25 | ST 0 | Frequency in MHz |

If Program A has not been used, only then,

46.96	ST 7	Z_1
37.86	ST 8	Z_2
-65.16	ST 9	Z_3

2. Follow loading instructions on reverse side
3. Press S/S. 1953.80 displays
4. | To Read | Press | Answers |
|---|---|---|
| Z_1 | RCL 4 | 5.98 in microhenries |
| Z_2 | RCL 5 | 4.82 in microhenries |
| Z_3 | RCL 6 | 1953.80 in picofarads |

Program Number 8807471-B Page 2 of 2

OPERATING INSTRUCTIONS

Loading the Program

1. Enter 1 ST 0
2. Switch to LOAD , PROG 2
3. Key in program
4. Switch to RUN
5. Set DP as desired

Running the Program

6. Enter and store the variables:

Enter	Press
f_o (MHz)	ST 0

If used without Program A, then

Z_1 (ohms)	ST 7
Z_2 (ohms)	ST 8
Z_3 (ohms)	ST 9

7. Press S/S Z_3 displays
8. | To Read | Press |
|---|---|
| Z_1 | RCL 4 |
| Z_2 | RCL 5 |
| Z_3 | RCL 6 |

REGISTER

0	1	2	3	4	5	6	7	8	9	F2
f_o				Z_1	Z_2	Z_3	Z_{1i}	Z_{2i}	Z_{3i}	

PROGRAM STEPS

1	RSET	41	
2	RCL	42	
3	0	43	
4	X	44	
5	PI	45	
6	X	46	
7	2	47	
8	=	48	
9	ST	49	
10	4	50	
11	X	51	
12	RCL	52	
13	9	53	
14	=	54	
15	1/X	55	
16	X	56	
17	1	57	
18	EXP	58	
19	6	59	
20	=	60	
21	CHSN	61	
22	ST	62	
23	6	63	
24	RCL	64	
25	8	65	
26	DIV	66	
27	RCL	67	
28	4	68	
29	=	69	
30	ST	70	
31	5	71	
32	RCL	72	
33	7	73	
34	DIV	74	
35	RCL	75	
36	4	76	
37	=	77	
38	ST	78	
39	4	79	
40	S/S	80	

fig. 2. Part B of T-network design program which uses the results of the program shown in fig. 1 to calculate the actual network inductance and capacitance values.

program 2. Examination of the procedure shows that you recalled the values in registers 7, 8, and 9 then calculated the component sizes to produce the new values and stored them into registers 4, 5 and 6, which had been used in program 1 but are no longer needed — a case in point to illustrate destructive replacement.

In considering the 3x3 portion of the keyboard of the 324G microcomputer, you have a complete grid under the nine registers. Using the recall button registers 1, 2, and 3 have the design param-

1, 2, or 3 we can enter or store the new value in the register, switch to program 1, hit START/STOP, then switch to program 2, hit S/S again, and you will have an entirely new and different set of network values. The program described applies not only to matching antenna impedances but may be used for intermediate or output stages of transmitters, delay lines, or any problem of a similar nature in impedance matching. The designer always has control of the parameters he enters, and he may wish to make a number of trials before

deciding on the most economically feasible set of final components.

You might ask, "What does all this buy me?" On the presumption that you've followed the previous paragraphs and understand the manipulation or game plan, it's only a jump upward into BASIC or FORTRAN IV. You've already learned an elementary form of assembly language.

final remarks

The average reader will find it's easy to generate programs within a very short time. There's really nothing complicated if a few simple ground rules are observed; i.e., start as deeply within the equation as possible, attacking the constants, and store for recall after the variables have been processed. As stated earlier, use the various registers for short storage, enter, kill, or substitute as you manipulate your way out. If you want to pause to read intermediate data, it's as easy as pressing the START/STOP button, but be wary of using this mode too often since you and the machine may become out of phase. When you want to bracket something, a set of parenthesis is available, which can be doubled if you want.

A little green man is inside the machine to count the number of entrance parentheses and make sure you use the same number of closed parentheses before you terminate the program. He's the same fellow who places the machine in lock-up if you violate any of the rules; but forgiveness is instant with the reset button, and you are permitted to start over again.

Programming and debugging come only with practice. There are all kinds of tricks that an operator will develop on any computer, whether using the simple program presented here or one on the big IBM-360 or IBM-370. The union of man and machine occurs only when one plus one functions as one!

ham radio

electronic bias switching for linear amplifiers

A simple electronic vox biasing circuit that lowers quiescent plate dissipation of linear power amplifiers

Marv Gonsior, W6VFR, 418 El Adobe Place, Fullerton, California 92635

The use and value of an electronic bias switch has been known for some time, having been introduced in the popular ETO Alpha 70/77 linears. Bryant subsequently published an excellent version of that circuit in *QST.*[1] It has significant application in two-kilowatt PEP input class-B and class-AB1 linear amplifiers where, especially in the latter, nominally two-thirds of the total plate dissipation is required in the quiescent state for best linearity.[2] This immediately gives rise to the impact of such a device as a means of reducing tube dissipation, the accompanying heat, the ambient noise, as well as lengthening tube life and saving some of your power bill. One further consideration is the possibility of reducing the fan speed, its resultant noise and power consumption. Also, the high class-AB1 idling current is no longer a problem, permitting greater linearity.

circuit

Basically, the bias switch data and background were well covered by Bryant, and served as an excellent basis for a modified design to fulfill my particular requirements, which I prefer to call *Electronic Vox Biasing*, or as a friend dubbed it, a *Vox Box*. That more accurately describes the action of my circuit shown in fig. 1. This was, I felt, necessary because the original circuits were much too fast, resulting in a very harsh vox-like action, especially on the make of the switch, and other operating time constants, which were far too short, giving rise to what might be described as a paper-crunching sound, especially at the end of a sentence during ssb operation. For CW break-in operation, however, it would be very effective.

The problem centered on how to slow down an inherently fast circuit due to the saturation characteristics of the transistors resulting from the high current gain of the Darlington configuration. After considerable work an integrator circuit, consisting of C3 and R3, was finally developed which allowed for a softer make that provided an almost indiscernable operation of the switch. The speed on the make, or ris

time of the switch, was changed by a factor of 20, from 2 to 40 milliseconds. The decay time of 140 milliseconds is no problem at all since C2 and R2 may be easily altered to any required time constant.

It should be recognized that the quiescent bias voltage itself at point A in **fig. 1** supplies V_{cc} to the Darlington pair. This is the self-bias developed by current, I_p, drawn by the cathode holds it in operation in its active mode. In my particular case, the voltage at point A varies from 55 volts to 0.7 volt (the $V_{ce(sat)}$). Depending upon the particular operating biases and the tube used in the linear amplifier, the quiescent bias voltage will vary. Since, for the reasons stated above, the switch will only operate when it is installed in the cathode

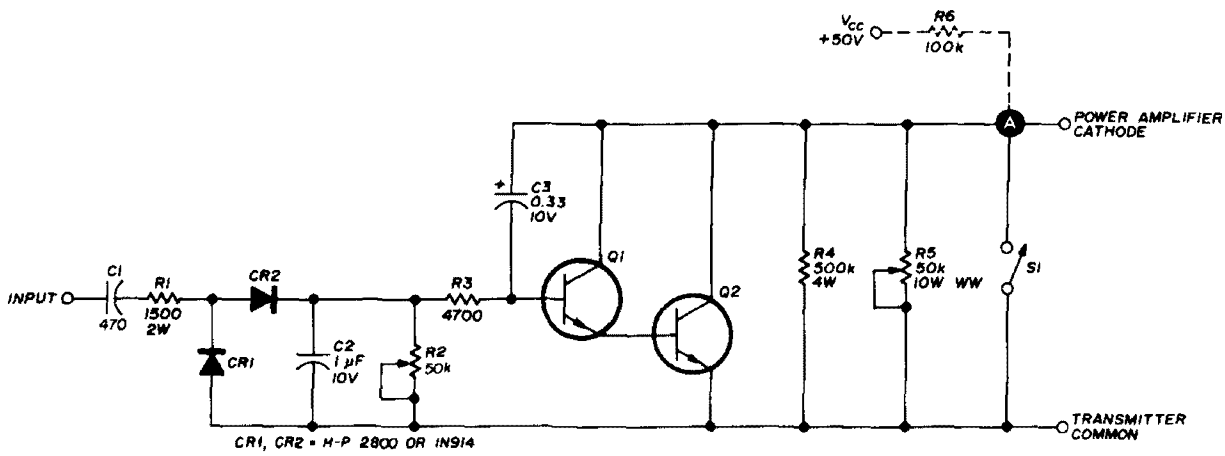

fig. 1. Electronic vox biasing circuit. The transistors used for Q1 and Q2 depend upon the cathode voltage of the power amplifier tube (see table 1). Adjust C1, R1 for drive level. Switch S1 permits establishing operating point for class-AB1 amplifiers and is used as operating switch in case of bias circuit failure.

the cathode current across resistor R5, which becomes ordinary center-tap bias on the amplifier. The system operation becomes apparent as an electronic normally-open spst relay which removes the unwanted self-bias in its closed state with appropriate time constants.

As has been shown, the Darlington pair has its V_{cc} supplied by the cathode voltage in its quiescent mode. The plate circuit, for independent test it is necessary to temporarily supply the V_{cc} separately through a 100k limiting resistor, R6, to point A.

table 1. Transistors for the electronic vox biasing circuit.

	output voltage ≤50 Vdc	output voltage 50-125 Vdc	output voltage 50-125 Vdc
Q1	2N5681 (TO-5)	2N3439 (TO-5)	2N3439 (TO-5)
		2N3440 (TO-5)	2N3440 (TO-5)
Q2*	2N5681 (TO-5)	2N3439 (TO-5)	2N6262 (TO-3)
		2N3440 (TO-5)	2N6354 (TO-3)
			RCA411 (TO-3)
			DTS423 (TO-3)

*Use heatsink with TO-5 devices

Table 1 outlines the various switching devices which will satisfy the requirements for Q1 and Q2 with V_{ceo} ratings greater than 100 volts. Further, a heatsink such as an *IERC* type *UPTO-3* may be added to transistor Q2 which *will considerably extend its 100-watt* power dissipation. The heatsink, used in combination with the necessary heat-conducting grease, will satisfy the most demanding requirement.

The serious designer could consider a total redesign of the circuit starting with an op-amp shaper driving a transistor as a voltage follower, rather than the Darlington configuration. This would provide complete control of the rise and

decay times. This will be something for a future project.

construction

I built my electronic vox biasing circuit on one-sided, copper-clad Vector board as shown in fig. 2. It only took a couple of hours to build, and could be done in less time if an unclad board were used since a good ground plane is apparently not required. Parts location is very noncritical and no problems were encountered with any instability, etcetera. Switch S1 may be a miniature relay, as I used, or a simple spdt toggle switch to short the bias switch to set the proper operating quiescent current, I_p, or to eliminate the switch in case of an open-type failure.

In the case of a shorting-type failure, the electronic vox biasing circuit will hold I_p to its normal switched value of 0.7 volt, i.e., essentially zero biasing. In the event resistor R5 were to open up, resistor R4 will prevent the full plate voltage from appearing on the cathode. These fail-safe precautions are in order for obvious reasons and will provide the user with a reasonable amount of voltage and dissipation security, since the total I_p flows through transistor Q2 along with the aforementioned high voltage risk.

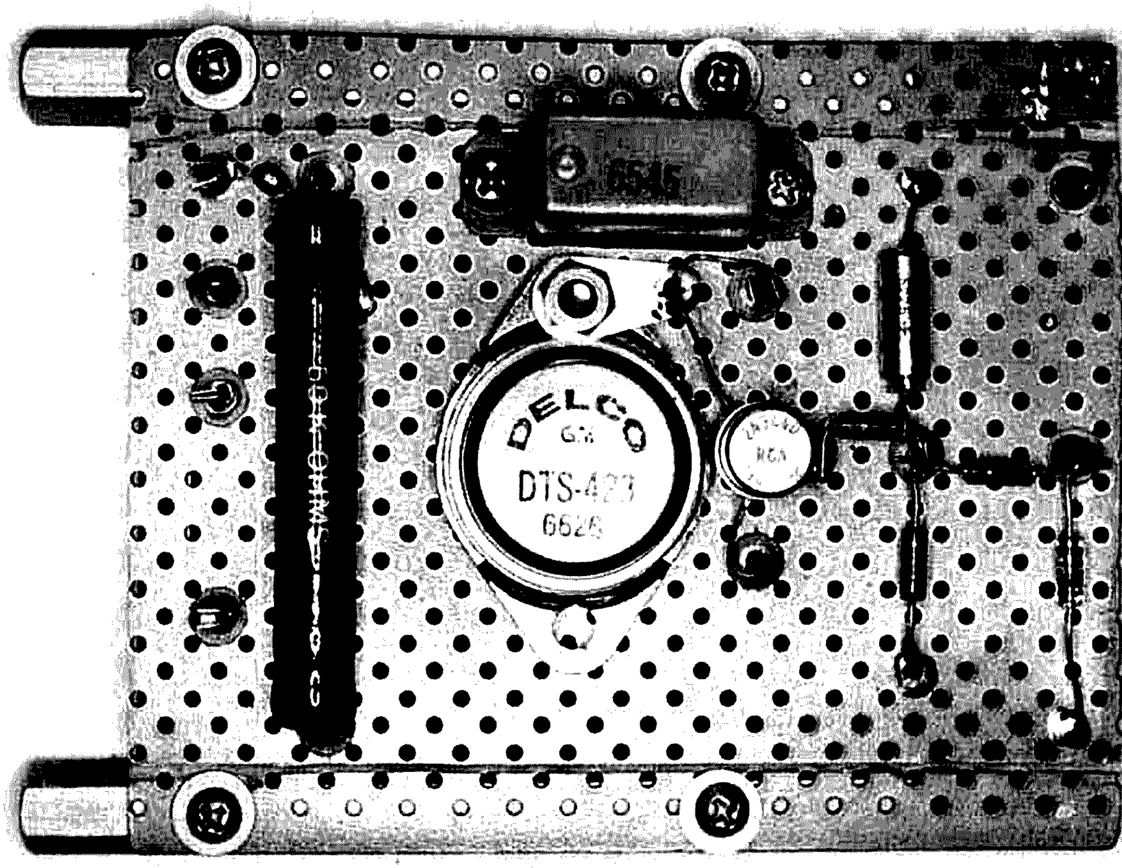

fig. 2. Construction of the electronic vox biasing circuit built by W6VFR. Although copper-clad perf board is shown here, a ground plane is not necessary, and since parts placement is noncritical point-to-point wiring is satisfactory.

Fig. 4 depicts the switch operation at a half-second sweep speed, with a male voice saying, "one, two, three, four." Fig. 5, at the same sweep speed, depicts the word, "four." As may be seen, the switching speed looks much like conventional CW keying time constants.

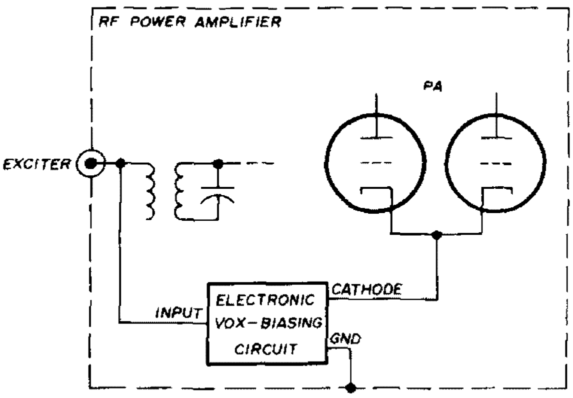

fig. 3. Typical application for the electronic bias-switching circuit. In rf power amplifiers using directly-heated tubes (such as the 3-500Z), the cathode output of the bias-switching circuit is connected to the center-tap of the filament transformer.

A few words are in order regarding the *time constants. Resistor R2, in* combination with C2, basically establishes the decay time of the switch with some interaction with the make time. Capacitor C3 in combination with R3 affects the make and the break, while the combination of R1 and C1 establishes the keying sensitivity and the hold time. These should be set for the reliable operation of the switch at the highest frequency of operation and at the lowest anticipated power level.

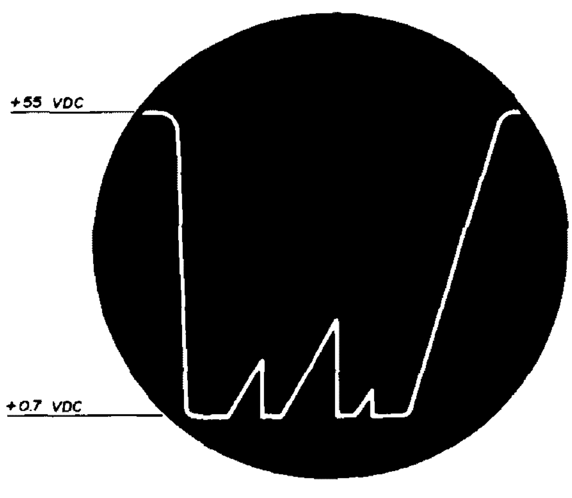

fig. 4. Oscilloscope display of electronic vox biasing circuit voltage output with male voice saying, "1, 2, 3, 4." Sweep speed is 0.5 second.

A dc-coupled scope is highly recommended for adjustment for all time constants, bearing in mind that the current gains of the semiconductors and variations in component values will alter the result to some degree. All of my timing measurements were made on a H-P 1220A oscilloscope. At 14 MHz, the keyer will reliably turn on completely at 0.5 volt rms as indicated on the output meter of a Measurements Corporation model 65B signal generator and verified with the oscilloscope.

The results of numerous on-the-air tests, both under local and skip conditions, indicate little or no perceptible switching action, and a 32°F temperature *reduction* in the outlet air temperature of *my linear amplifier.* This, at 55 cfm air flow, calculates to a 520 watt *average* power reduction, a well worthwhile improvement by any criteria.

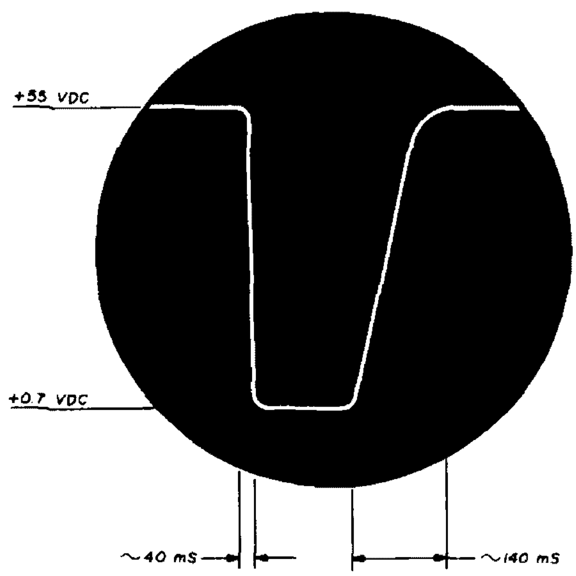

fig. 5. Oscilloscope display of electronic vox biasing circuit voltage output with male voice saying, "four." Sweep speed is 0.5 second.

references

1. J.A. Bryant, W4UX, "Electronic Bias Switching for RF Power Amplifiers," *QST*, May, 1974, page 36.
2. *Linear Amplifier and SSB Service Bulletin 12*, Eimac, San Carlos, California, 1966.

ham radio

low-power dc-dc converter

A simple transformerless dc-dc converter circuit which may be used for those applications where you need a separate low-current supply

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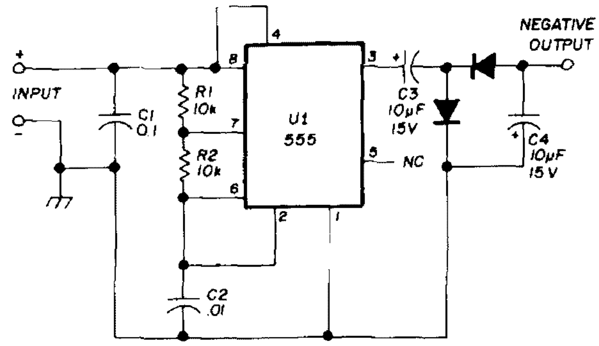
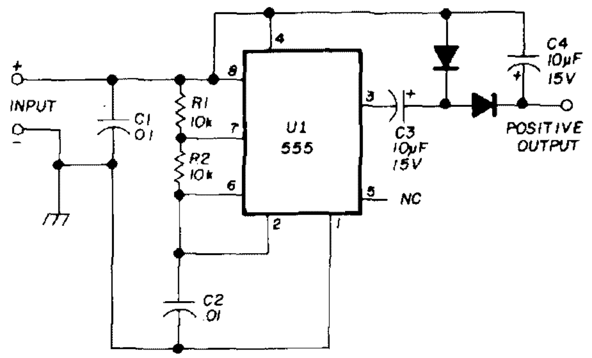
When designing electronic systems around ac power, an additional supply voltage seldom represents any problem. For low power an additional rectifier and/or regulator will often suffice, and worst case means the addition of a small transformer. When mobile operation is considered the problem becomes more difficult. If the requirement is for a voltage of the same polarity and lower than that of the automobile battery, then a simple voltage regulator will do the job. However, if you need a voltage that is higher or of opposite polarity than the vehicle battery, then something more elaborate is indicated. Occasionally it is possible to obtain power from an existing converter and suitably regulate it, but with more and more equipment being designed for direct 12-volt operation this option is fast disappearing.

The transformerless dc-dc converter described here may be adapted to a variety of low-power applications requiring voltages either higher or of opposite polarity than the vehicle battery. This power supply may also find some application as an on-card supply in fixed installations. As a negative voltage supply it may be used to power linear ICs such as operational amplifiers. By



A



B

fig. 1. Low-power dc-dc converter may be used to supply negative voltages (A) or positive voltages of greater value than the input (B). For positive-ground systems, allow the ground shown in these circuits to float. Voltage-current capabilities of these circuits are listed in table 1. All diodes are 1N914, 1N4148 or equivalent.

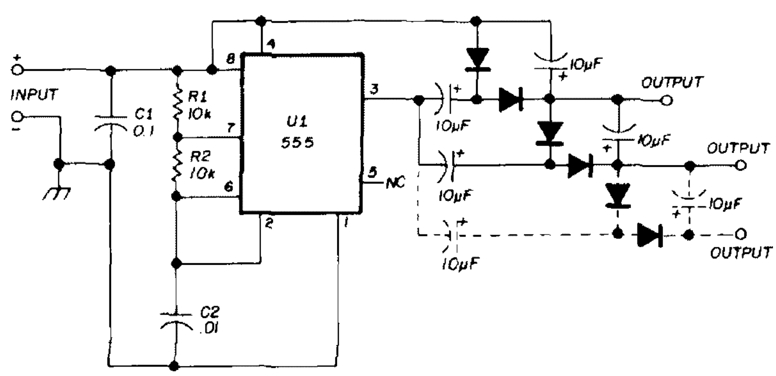
using the vehicle battery as the positive supply and the dc-dc converter for the negative supply, signals may be referenced to the vehicle ground. This is very convenient, especially in control applications. As a voltage booster supplying voltage of the same polarity but of a higher potential it provides a convenient source for such things as trickle charging 12-volt nickel-cadmium batteries.

The basic negative converter is shown in **fig. 1A** and the positive booster shown in **fig. 1B**. **Fig. 2** shows how additional diode-capacitor voltage-doubler sections may be added to increase the voltage output of the positive voltage booster (at the expense of available current). The same principle may also be applied to the circuit of **fig. 1A** to produce negative voltages of higher potential. Although all the schematics indicate a negative-ground input power source, positive-ground input may be used by grounding the positive input to U1 and floating the ground shown in **fig. 1** and **2**.

operation

Operation of the dc-dc converter is straight forward. U1 is a 555 timer IC operated as a free-running square-wave oscillator. The 555 makes an ideal oscillator for this application as it requires only three external components for oscillation and the output has the ability to source and sink up to 200 mA without additional buffering. The frequency of operation is determined by R1, R2 and C2 (with the values shown it is approximately 6 kHz). Capacitor C1 is used to reduce the amount of

fig. 2. Voltage output from the positive voltage booster (fig. 1B) may be increased by the use of voltage multipliers as shown here. Diodes are 1N914, 1N4148 or equivalent.



6-kHz signal *radiated* through the input power lines.

Since the switching times of the 555 IC are quite fast, this 6-kHz signal generates harmonics up into the higher high-frequency bands and may be heard on a communications receiver if the switching transients are not filtered. If the dc-dc converter is operated as part of a high-frequency receiver, it may be necessary to *insert small (100-μH) rf* chokes in the power leads to further reduce these harmonics.

The 6-kHz output at pin 3 of U1 is capacitive coupled by capacitor C3 to a rectifier circuit and filtered. The output filter capacitor, C4, may be increased above the 10 μF value shown if a low ripple content is essential. However, for operating an op-amp or two the amount of ripple with the circuit as shown is not objectional for most simple applications. Voltage-current capabilities for various voltage inputs are listed in **table 1**.

table 1. Voltage-current capabilities of the dc-dc converter shown in fig. 1. For circuit of fig. 1B, add input voltage to output voltage (i.e., 5-volt input with no load, E output = 5 + 3.8 = 8.8 volts).

Input = 5 Vdc	Input = 6 Vdc	Input = 9 Vdc	Input = 12 Vdc
output	output	output	output
3.8V, 0 mA	4.8V, 0 mA	7.7V, 0 mA	8.9V, 2 mA
2.2V, 1 mA	3.2V, 2 mA	6.2V, 1 mA	8.4V, 10 mA
2.1V, 2 mA	3.0V, 4 mA	6.1V, 2 mA	8.1V, 15 mA
2.0V, 3 mA	2.9V, 6 mA	6.0V, 4 mA	7.9V, 20 mA
1.9V, 4 mA	2.7V, 8 mA	5.9V, 6 mA	6.7V, 30 mA
1.9V, 5 mA	1.5V, 10 mA	5.7V, 8 mA	6.1V, 40 mA
1.8V, 6 mA		5.6V, 10 mA	5.7V, 50 mA
0.6V, 7 mA		5.3V, 15 mA	5.5V, 60 mA
		5.0V, 18 mA	5.0V, 70 mA
		4.2V, 20 mA	4.6V, 80 mA*
			4.0V, 90 mA*

*Current exceeds rating of recommended diodes.

construction

Construction of the dc-dc converter is not critical although capacitor C1 should be as close as possible to U1 to reduce radiation of the 6-kHz harmonics. As it is anticipated that the converter will probably be incorporated as part of another circuit board, no PC layout is shown. For prototype applications, however, it may be convenient to build up a few on small PC boards (the supply shown in **fig. 1A** will fit with room to spare on a PC board 3/4-inch wide by 1½-inch long (19x38mm).

ham radio

"I wish I could establish a single transmission for almost three minutes with you."

brass pounding on wheels

The saga of Old 1401 — first radio communications from the rolling White House, the presidential train

Charles W. Clemens, Jr., K6OD, 9971 Deerhaven Drive, Santa Ana, California 92705

Many of the operating stories we read in the amateur publications are about sea-going brasspounders. This one is about a brass pounder who rode the rails. Back in the summer of 1942, I was working my shift at WAR in Washington when an officer walked up behind me and tapped me on the shoulder. He told me to go pack my clothes for a trip to a warm climate. That was how I started as the first CW operator at the White House.

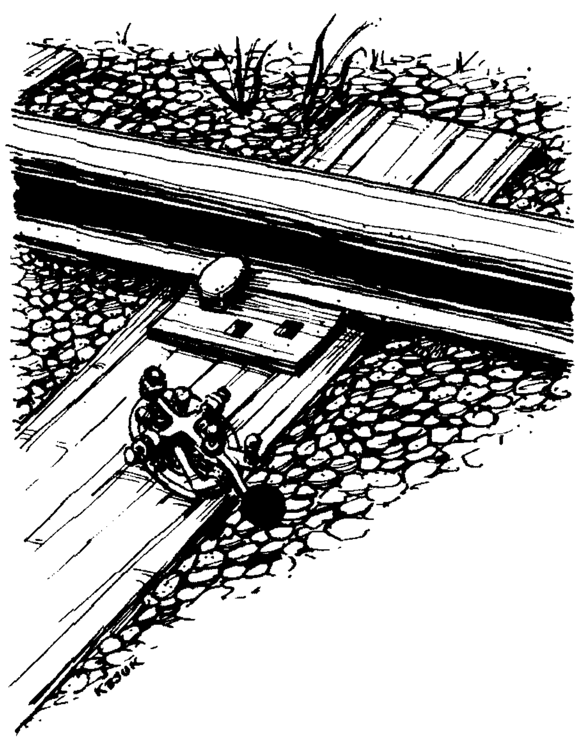
I learned that the White House had a Signal Corps detachment that now had the task of providing communications on a continuous basis between the Presidential Train and the White House. I *believe this was the first time such a thing* had been attempted in the United States. The Washington end was to be handled by the big War Department communications center, WAR; the remote end by the train plus relays, when necessary, from local stations along the way.

My first trip on the communications car, *Old 1401,* was the second trial run for the car. On this trip I went with the detachment commander, Col. Beasley; a radio operator recently made a Lieutenant, Lt. Greer; a civilian engineer from the War Department named Jack Kelleher; a radio maintenance man named John J. Moran; and a Secret Service man named George J. McNally. It is interesting to note that all of us were amateur radio operators. We went from Washington, D.C. to New Orleans and returned with our car in a regular passenger train, coupled between two baggage cars.

Old 1401 was a *combine car.* That is, she was half baggage and half passenger. She had been built for the Baltimore and Ohio Railroad in 1914. At the time I first met her, all identification on her sides had been painted out. Her number was her only identification, and it was painted in beautiful gilt over the entrance at the passenger end.

Inside, a couple of front seats had been removed and an operating table installed in their place. One operating position was located on each side of the aisle between the seats. Each position had a Super Pro receiver and a BC-342.

The BC-342 was a new model at that time, designed for use in tanks and other rough riding vehicles. This receiver was installed on shock mounts, but my first trip *proved that the best way to mount* equipment on the train was to bolt it down solidly. Installed in this manner, the whole car moved as one unit and the



receivers worked beautifully. There was, however, a modulation on the received signals imparted by the train's vibration. But this was better than having the tubes jump out of their sockets — which they frequently *did when* the equipment was on shock mounts (the tube clamp was not yet in common use).

Telegraph lines alongside the tracks provided a lot of clicks that made it difficult to copy poor signals. However, we didn't have too much trouble with this problem except in the Southwest. The transmitter was a BC-447, running about 300 watts.

The clearance requirements for railroad cars prohibited using a real antenna. Ours was a wire inside an insulating tube mounted on standoffs about six inches above the metal roof of the car. This was later changed to a copper tube, the same size as the insulating tube, with much better results. Our frequency complement ran from 3 MHz to 17 MHz.

I was supposed to contact a number of Army stations along the way, none of them more than a couple of hundred miles from our route. As might be expected, results were poor and it was decided to contact WAR in Washington direct. Successful contacts were made from New Orleans and on the way home. The only real difficulty came when we were close to Washington. At that time, it was difficult to receive WAR on any frequency. Overall, however, our results were encouraging and we were assigned the task of accompanying President Roosevelt on his swing around the country visiting military bases and aircraft plants.

To my knowledge, this big trip was the first time continuous communications had ever been attempted between the presidential train and Washington. We contacted WAR in the eastern half of the country and WVY (San Francisco) or WVD (Seattle) in the western half. Results were excellent. In fact, our volume of traffic was so high that it was necessary to pick up an additional message clerk in Seattle, our first major stop, to handle the paper work.

To make a long story quite short, I worked six years on the Presidential Train, traveling with Presidents Roosevelt and Truman in the United States, Canada and Mexico. We logged well over a hundred thousand miles.

Equipment and facilities were improved over those years, and when I left *Old 1401* in 1948, the car had a small operating room, a code center, a small bunk room with four bunks, a lounge room, and the baggage half of the car packed with equipment. We had two BC-339 transmitters for our message traffic. These were fixed station Federal jobs that loafed along at 1500 watts in radioteletype service and could easily run 3 kW on CW.

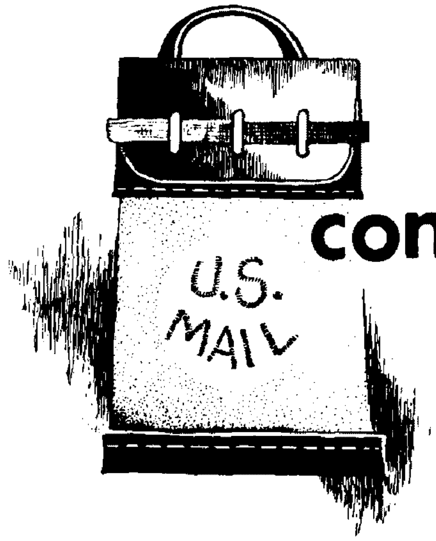
A single BC-610, a 500 watt a-m transmitter, was available for occasional broadcast services. We also had a 250-watt Motorola fm transmitter for guard radio service. On the receive side, we had the two BC342s I mentioned earlier, two Super Pros, a big Navy receiver whose type number I can't recall, two Western Electric CV-31 teletype converters and a single teleprinter.

We also had a telephone switchboard that provided service throughout the train. The telephone cable permitted us to provide music throughout the train and intercom service, too, if it were desired.

Power was provided by two 25-kW diesel generators. Only one of these was required, and we switched them every 24 hours. We also had two 100-amp battery chargers to charge the train's batteries when we were parked away from railroad terminal facilities, and two converters to provide ac power from the batteries to run our receivers in standby.

Today, the train is no more. *Old 1401* has been retired and the President's car — *Ferdinand Magellan — is gone, too.* The small detachment I knew has grown to the White House Communications Agency. Their responsibilities have grown a great many times over. But I'll bet they aren't having any more fun working assignments today than I did when *Old 1401* was my home on wheels.

ham radio



comments

impedance bridges

Dear HR:

I noted the two items on noise bridges in the May, 1974, issue and perhaps my experiences with the two bridges might be useful. First, regarding WB2EGZ's bridge (December, 1970) I had the same problem, i.e. insufficient gain. I solved this by the use of two stages of amplification using the high beta RCA transistor 40245 before going into the balun transformer. This yields a very simple circuit and only a 5 mA drain on the battery, an important point because when making measurements the unit is often run for long periods. Since this bridge measures only the parallel equivalent resistance of the load, the presence of a significant amount of reactance greatly reduces the depth of the null.

The bridge design presented in the January, 1973, issue does measure reactance and consequently the depth of nulls within its range of reactance tuning can be as deep as those obtained with pure resistances. However, I found that the plus/minus 70-pF range for reactance to be inadequate. An examination of the possible capacitive or inductive components encountered around a 2:1 vswr circle on a 50-ohm Smith chart indicates, for example, that if R_{series} = 58.8 ohms and X_{series} = ±38 ohms, the parallel equivalents are R_p = 83.3 ohms and X_p = 129 ohms. This translates into a Cp of ±324 pF at 3.8 MHz or ±88.2 pF at 14 MHz. The consequence is that a bridge capable of reading ±70 pF can only yield information about an antenna on these bands whose vswr is already pretty good. I changed this bridge to use a 365-pF variable (Radio Shack has a small one for $1.95). The 68-pF fixed capacitor then has to be changed to 180 pF. Furthermore, I brought out terminals from these capacitors so that I can add further fixed capacitors as needed.

One point needs emphasis when working with these bridges and Smith charts: these particular bridges read out the parallel equivalents of resistance and reactance, while the Smith chart is designed for the series equivalents. The following series-parallel transformation equations should be kept handy:

$$R_s = \frac{R_p X_p^2}{R_p^2 + X_p^2} \qquad X_s = \frac{R_p^2 X_p}{R_p^2 + X_p^2}$$

$$R_p = R_s + \frac{X_s^2}{R_s} \qquad X_p = X_s + \frac{R_s^2}{X_s}$$

It can also be shown that:

$$\frac{R_p}{X_p} = \frac{X_s}{R_s} = Q$$

One further comment: While the R_p calibration holds pretty well from 3.5 to 30 MHz (paying careful attention to strays), the C_p calibration rotates towards the inductive direction by about 20 pF at 30 MHz for R_p values between 35 and 200 ohms, more so for values under 35 ohms. If the experimenter is using this bridge at 30 MHz or higher it is important that he calibrate the bridge for these frequencies. I found that a calibration made at 3.5 MHz holds pretty well through the 21-MHz band.

Provided one knows the characteristics of the coax line (characteristic impedance, velocity factor, attenuation and electrical length) with this bridge and a Smith chart one can measure the characteristics of an antenna at the shack coax terminal. It is not necessary that the line be an exact half-wave multiple. The assumption that the load characteristics are faithfully represented at the end of a multiple of a half-wavelength transmission line is true only at the exact frequency for which the line is a half-wave multiple (see April, 1974, *QST* article by W2DU). Usually we are interested in what an antenna is doing over part or all of an amateur band and it is important to realize that for frequencies other than the exact half-wave multiple the load is no longer faithfully represented. The error becomes increasingly significant with increasing vswr.

Forrest Gehrke, K2BT
Mt. Lakes, New Jersey

ac current monitor

Dear HR:

The *ham notebook* section of the January, 1974, issue described a line voltage monitor that is very similar to one I have been using for about four years. The differences are slight, but my circuit has fewer components than WA8VFK's version. I didn't feel that the bridge rectifier and filter capacitor were required. If you are willing to accept only a little more non-linearity, a half-wave rectifier with bypass capacitors connected across the diodes and the input are all that is required. Rather than one expensive zener diode, I used three zeners from my junk box which added up to the voltage I needed.

A variable transformer in the ac power to my bench is used to check the performance of various types of equipment for over and under-voltage, and to compensate for line-voltage variations.

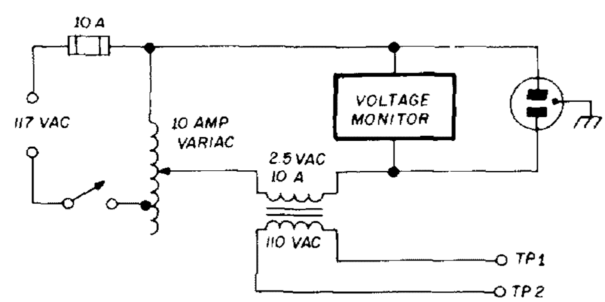


fig. 1. Ac current monitor uses high-current 2.5-volt filament transformer. One amp through 2.5-volt winding yields 11-Vac between TP1 and TP2.

A 2.5-volt, 10-amp transformer in the circuit (see **fig. 1**) is used to provide a readout of ac current on a voltmeter connected across TP1 and TP2. Each ampere of ac current passing through the 2.5-volt winding develops 0.25 volt across the winding — this is transformed to 11 Vac at TP1 and TP2 (1 amp = 11 Vac, 2 amps = 22 Vac, etc.). Any low-voltage, high-current transformer can be used in this application, but the 2.5-volt,10-amp unit is ideal because the transformed voltage is easily interpolated into current (1 volt ≈ 100 mA ac current).

E.G. Sullivant, Jr., WB5MAP
Shreveport, Louisiana

the ham notebook

increased flexibility for the memory keyer

In a recent article WB9FHC described the construction of a two memory electronic keyer.* After using this unit on the air for several weeks, I decided to expand the keyer's capability, permitting greater operating flexibility during contests.

As a first step, I wanted a self-contained keying monitor, particularly for initial off-the-air programming. A simple but effective circuit is shown in **fig. 1**. The circuit uses half of a 7413 dual NAND Schmitt trigger. A miniature 500-ohm pot is used for the tone control, while the inexpensive 8-ohm speaker was salvaged from an old transistor radio. The monitor can be switched out of the circuit when not in use.

Many times during a contest, especially during low activity periods, a message, such as a CQ, is to be repeated a number of times without having to manually recycle the memory each time. A simple solution is to install a spst toggle switch in parallel with S1 of the original circuit. In my particular case, a CQ without the $\overline{AR}$ K ending was desired to nearly fill the 256-bit memory capacity. During programming a stop-watch was used to note the elapsed time of the memory cycle. By careful spacing of the message CQ CQ DX TEST DE K3NEZ K3NEZ, the entire memory can then be filled without any appreciable pause at the end. If the message is to be sent three times, the toggle switch is closed and then opened just after the desired message starts to repeat for the third time. The $\overline{AR}$ K ending is then sent manually with the paddle after the callsign has been sent by the memory.

Since it was desired to add a third, and possibly fourth, 256-bit memory to the keyer, it became apparent that the use of a single rotary switch for selecting the memory was not efficient. The circuit of **fig. 2**, when added to the original circuit, permits the operator to simultaneously select the desired memory and cycle the four-bit counters.

Although three memory units are shown, additional units are easily paralleled, as the circuit is straightforward. A LED indicator is provided for each message. Note that the part of the original circuit from $\overline{Q}$ of 7473C is omitted. For example, to read or write from memory A, simply depress the corresponding switch; otherwise, operation is the same as described in the original article.

Howard M. Berlin, K3NEZ

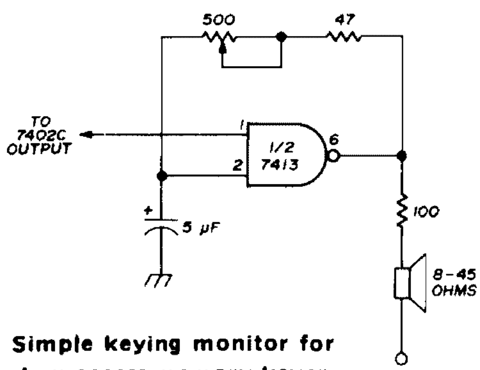


fig. 1. Simple keying monitor for the random-access memory keyer.

*Michael Gordon, WB9FHC, "Electronic Keyer with Random-Access Memory," *ham radio*, October, 1973, page 6.

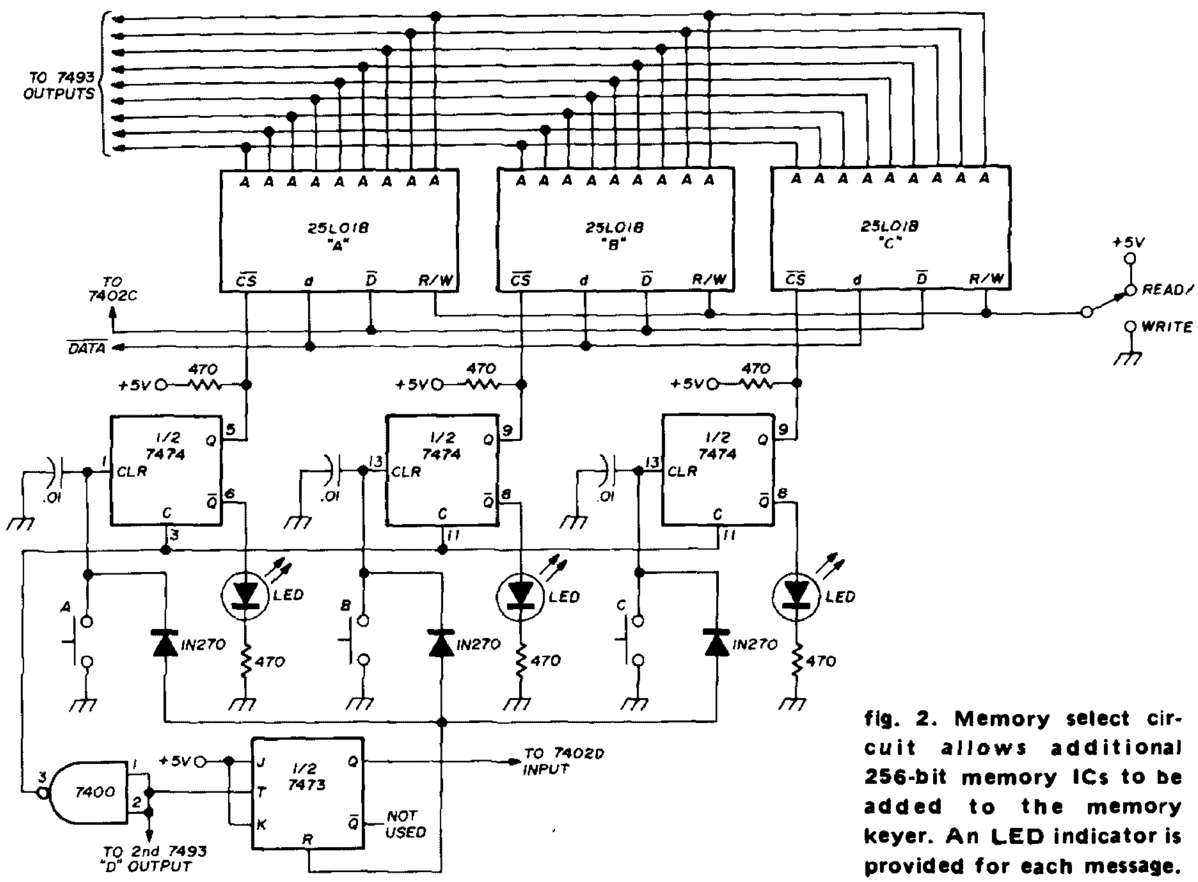


fig. 2. Memory select circuit allows additional 256-bit memory ICs to be added to the memory keyer. An LED indicator is provided for each message.

hi-fi interference

An excellent article on audio-frequency interference in a recent issue of *Radio Communications* discussed problems which are not mentioned in our usual Handbooks.

A simple LC filter was first tried on the bench to keep rf out of the coaxial lead to the audio amplifier, particularly from a record player. The coil turned out to be resonant, so it was not reliable at widely varying frequencies. The capacitor alone, to ground, was usually better. When tried in a Garrard record player, however, which has unshielded pick-up leads connected to phono jacks, *no improvement* was noted.

The rash of hi-fi complaints here results from unidentified CB as well as my own transmissions. The greatest problems were from Garrard players, but to a lesser extent from a Wollensak-3M cartridge player. When all input and output coaxial lines were removed the fm broadcast receiver played without interference, but one a-m tuner had many signal peaks in its tuning range.

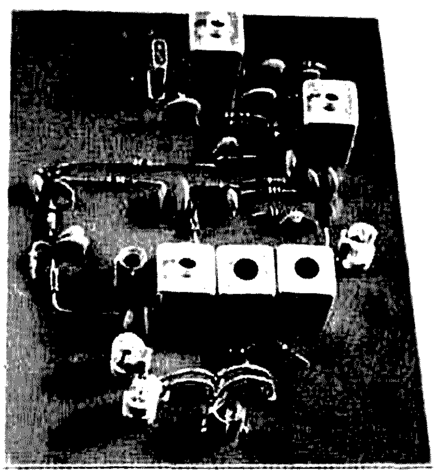
In addition to bypassing audio-amplifier base-emitter junctions with rather large capacitors (1000 pF up), I hope to try the recommended ferrite beads, and to develop lowpass filters to plug into the coaxial input and output lines without creating unacceptable loss of high-frequency audio.

Radio Shack tells me that several manufacturers *have* produced *filters* which can be inserted in the audio leads from record-players and tape-players to prevent rf pick-up from reaching the audio transistors in the amplifier. These filters are seldom known even to the hi-fi departments of stores, but are available upon request, much as high-pass TVI filters have been available to those who know about them and are familiar with this method of complying with the FCC regulations.

Bill Conklin, K6KA

new products

uhf converter and preamplifier



Hamtronics, Inc., well known for its vhf preamps, receivers and scanners, is moving into the uhf field with two new products. The first is a uhf converter kit, shown above, which operates on the 432- to 450-MHz amateur band or on the rapidly growing 450-470 MHz public safety band. The converter is constructed on a 3x4-inch G-10 board and features a low-noise jfet, high-quality, milled variable capacitors, integral coax connectors, and low current 12-volt operation.

The converter has one built-in oscillaor, and an adapter is available for six additional frequencies for channelized fm operation. It can be built to output on the popular two-meter fm band or 6 meters, 10 meters, commercial bands, etc. Price for the kit is $20, which includes domestic shipping. Add 80¢ for air mail if desired. Crystals are available for any frequency scheme at $5.50.

The second new uhf product is a low-noise preamp kit for 432-450-470 MHz. This unit has features similar to those of the converter. Price is $15 for the kit or $25 wired, including domestic shipping. Add 60¢ for air mail if desired. For more information or to order, write to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

how to troubleshoot and repair test equipment

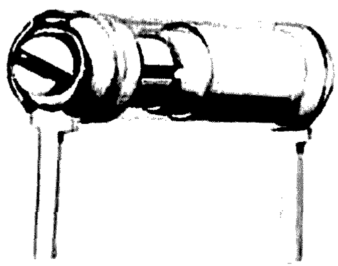
"Cure your own equipment" is the watchword of this brand-new volume, written by Mannie Horowitz, one of the top designers of electronic test equipment. Any competent amateur should be able to repair his own equipment, and it's easier than ever before with this new one-of-kind book! It's jam-packed with practical, ready-to-use data on the repair of power supplies, multimeters, oscilloscopes, audio and rf signal generators, sweep generators and tube and semiconductor testers. No complex math or circuit theory is included — this is *not* a book that requires study, but one that can be used from the very minute it's opened. With the complete, simplified theory that is presented, read-

More Details? CHECK—OFF Page 94

ers with even modest electronic backgrounds will understand equipment as never before, and get more out of using it as well.

For each piece of equipment there is a clear, illustrated explanation of the basic circuits it contains. There's also a complete trouble analysis of each circuit, telling what can go wrong and what the probability of it is. Next comes an explanation of how the basic circuits are integrated into a complete circuit by the switching circuitry. Then, test procedures for an actual example are presented. Published by Tab Books, 252 pages, soft bound, $6.95 from Ham Radio Books, Greenville, New Hampshire 03048.

high-precision trimmer capacitor



A new type of miniature trimmer capacitor claimed to give outstandingly linear response (better than 2%, with no local reversals of capacitance) has been announced by Jackson Brothers of England. The *Trimline* capacitor is a tubular design 5-mm in diameter and 18-mm long. Its contant length simplifies layout planning. Minimum capacitance is below 0.5 pF and maximum is above 5 pF. Adjustment is by screwdriver slot, with ten turns between minimum and maximum to permit very fine setting.

Unlike most tubular trimmers, the *Trimline* employs air as the dielectric, and the moving element does not rotate. This avoids eccentricity and deviations

from linear response associated with designs based on a rotating piston. Here the stationary element is a small piston and the moving element is a coaxial cylindrical sleeve. Both are made of silver-plated brass. The sleeve is moved axially by a lead-screw engaging a threaded collet inside it; it is precisely located and guided by a fixed outer glass sleeve, and by two lugs in its tail which run in fixed longitudinal slots.

At both ends of the travel a slipping-clutch mechanism — a patented feature — prevents accidental damage being caused by over-adjustment. At maximum capacitance a Teflon stop prevents electrical shorting. Electrical losses are low (Q factor greater than 1000 at 20 MHz), and the capacitor has an expected life of over 10,000 adjustment cycles. For more information, write to Jackson Brothers' agent, M. Swedgal, 258 Broadway, New York, New York 10007, or use *check-off* on page 94.

tool catalog

A free tool catalog describing over 2500 individual items is offered by Jensen Tools and Alloys. "Tools for Electronic Assembly and Precision Mechanics" is a 112-page handbook of particular interest to electronic technicians, radio amateurs and engineers.

Section headings include screwdrivers, wrenches, pliers, tweezers, files, shears, knives, microtools, relay tools, power tools, metalworking tools, wire strippers, soldering equipment, test equipment, engineering and drafting supplies and electronic chemicals. New sections include metric tools, books and wire-wrapping tools.

Another important feature of the catalog is the inclusion of four pages of technical data on tool selection. Known as "Jensen Tool Tips," these pages include sections on screwdriver selection, machine screw data, tool materials,

More Details? CHECK—OFF Page 94

plier facts, metal conductivity, color coding, wire and insulation data, solderability of metals, temperature conversion, drill sizes, metal gauges, metric conversion and safety. Five pages of "tool terms" are also included.

A free copy of the Jensen catalog may be obtained by writing to Jensen Tools and Alloys, 4117 North 44th Street, Phoenix, Arizona 85018, or by using *check-off* on page 94.

diode applications

Diodes are the *simplest* yet *most* versatile devices found in electronic circuits. Typical applications range from power supplies to waveform converters, logic elements, temperature-compensating devices, regulators and signal detectors. This new book by Courtney Hall, WA5SNZ, surveys these and many other important uses for diodes.

In a clear, readable presentation, the book begins with basic information about diode properties and how diodes work. *Vacuum-tube diodes and* rectifiers, semiconductor diodes and diode reverse recovery time are other topics covered in Chapter 1. Chapter 2 discusses rectifier power-supply circuits. Half-wave power supplies, full-wave rectifier circuits, voltage-doubler circuits, transformer ratings and more are explained in depth.

The next chapter covers circuits for ordinary diodes. Beginning with logical OR and AND circuits, it covers the ideal diode, diode gate circuits, flip-flop preset *circuits and diodes for* meter protection. Another chapter is devoted to zener diodes. Two chapters are set aside for the newer diode types such as LEDs, tunnel diodes, varactors and semiconductor lasers.

Soft cover, 96 pages, $3.50 from HR Books, Greenville, New Hampshire 03048.

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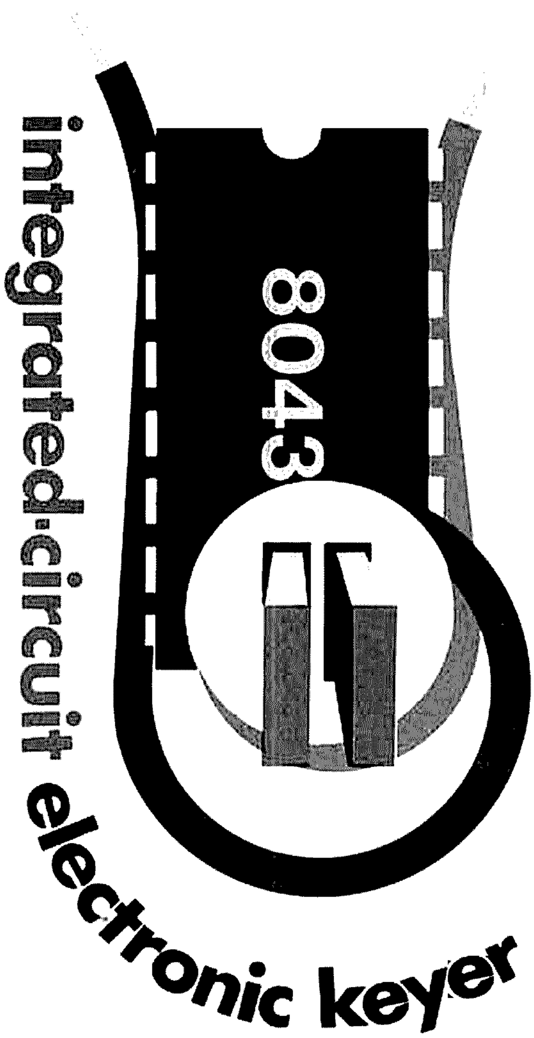
focus on communications technology . . .

ham radio magazine

APRIL 1975



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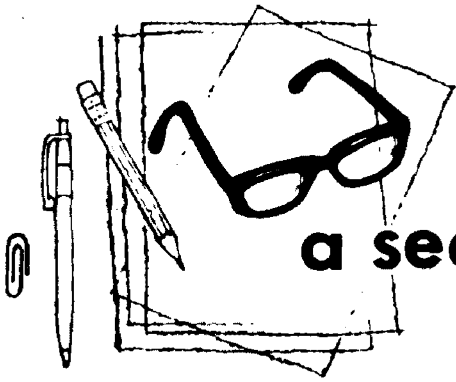


a second look

by jim fisk

During the past few months more and more vhf stations have been getting set up for serious moonbounce work. A good share of this activity has been in progress for months, but the February EME tests by WA6LET with SRI's big 150-foot dish in Palo Alto was the stimulant many vhf operators needed to become fully operational. Some operators, particularly in the northern areas, were hampered by the winter weather, but the seed has been planted so the number of vhf EME stations will probably double by the end of the summer.

Five years ago there were but a handful of successful EME stations – only the most serious and perservering workers ever made the grade. Those who did make it developed the techniques that are used today. And today equipment is available so that just about anyone who is interested can hear his own echoes off the lunar surface.

Right after World War II, when the first radio signals were bounced off the moon by engineers at Fort Monmouth, it was not nearly so simple: 8 kW input on 111 MHz, a tremendous billboard array of 64 phased dipoles and an incredibly complex receiver with a 50-Hz passband. Even then, the moon echoes were weak, and success unpredictable.

With this kind of background, the prospects of amateur communications via the moon were pretty remote, but W3KGP and W4AO launched Project Moonbeam in the late 1940s with a goal of hearing their own two-meter echoes from the moon. They finally heard them in 1953.

Until parametric amplifiers became activity was confined to two meters. EME attempts on 432 MHz were impractical because of the power limitation that was in effect, so the next logical step was 1296 MHz. In 1960 W1BU came on the air with a kilowatt 1296-MHz station that could bounce signals off the moon with some degree of reliability. Sam Harris, W1FZJ, extended the challenge and Hank Brown, W6HB, picked it up. Shortly thereafter the first *two-way* amateur contact via the moon's surface was history.

Shortly thereafter the first *two-way* amateur contact via the moon's surface was history.

Since 1960 progress has been steady if slow. Although all the vhf bands from 50 to 2300 MHz have been used for two-way EME contacts, at the present time most of the activity is found on 144 and 432 MHz. The handful of stations on two-meter EME a few years ago has grown to nearly 100, and there are about 35 stations operational on 432 MHz. Nor is activity confined to the United States. There are successful EME stations on every continent except Asia, and serious interest has been expressed by several Japanese amateurs, so Asia may be represented in the near future, possibly this summer.

EME is still a very sophisticated method of vhf communications, but with the easy availability of low-noise, solid-state converters and high-efficiency kilowatt amplifiers, it is within the grasp of any serious vhfer. Most of the successful moonbouncers live on city-size lots, so space is not a problem, and many use all commercial equipment, so gear is no problem. Most vhf stations capable of long-distance tropo, scatter or auroral communications, in fact, are also capable of two-way EME. It has taken twenty-five years, but the 1950's pie-in-the-sky dream of reliable two-way EME communications has become a reality.

Jim Fisk, W1DTY
editor-in-chief

presstop

DECISION ON 220-MHz CLASS-E CB is likely to be put off again. We had hoped for delay until after both amateur and CB restructuring go into effect and their results have a chance to be felt -- we'll probably get a delay that will permit the FCC to study all three dockets together. Any delay we do get will give us a chance to further strengthen our position, increasing the likelihood of preserving 220-225 intact for the amateur service.

WA6LET MOONBOUNCE TESTS brought many vhf stations out of the woodwork, made 53 two-way EME contacts on 144 MHz with 36 different stations during eleven hours of operating time. Score: 14 states and 7 countries on two meters. Most of the QSOs were on CW though five stations made it on ssb. In addition to stations in Germany, France, Sweden and the Netherlands, they worked ZE1DX in Rhodesia for the first U.S.-to-Africa EME contact.

432 MHz EME Test was shortened to two hours due to problems with antenna relays and an arcing HV power supply, but the WA6LET operators still managed to work ten stations via the moon on 432 including PAØSSB and ZE5JJ. K2UYH was the only station to make it both ways on ssb. More 432-MHz EME tests are expected later this year from WA6LET.

AUTOMATIC TRANSMITTER IDENTIFICATION SYSTEM is proposed by the FCC in Docket 20351 released February 13. ATIS requirement would apply to all transmitters licensed in Safety and Special Services produced after one year after adoption and operating between 25 and 960 MHz, with the exception of amateur radio.

ATIS Signal would be ASCII and give the station call sign at the beginning, end and every 30 seconds during a transmission. First or second class license would be required to program the encoder.

AMSAT REQUESTS U.S. DXER HELP in setting up interested DX stations for OSCAR 7, Mode B (432 MHz in, 144 MHz out) operation. AMSAT has acquired a number of uhf transmitters and has them tuned up and ready to go on 432 MHz, will provide one on a loan basis to anyone willing to take the responsibility of getting it into the hands of an overseas amateur interested enough to put it on the air. The DX station would still have to supply his own antenna and two-meter receiver.

OSCAR 7's Mode B has already seen some exciting DX -- both ZL and UAØ have been reported worked from the midwest, and WB6NMT reports hearing over-the-horizon signals several times.

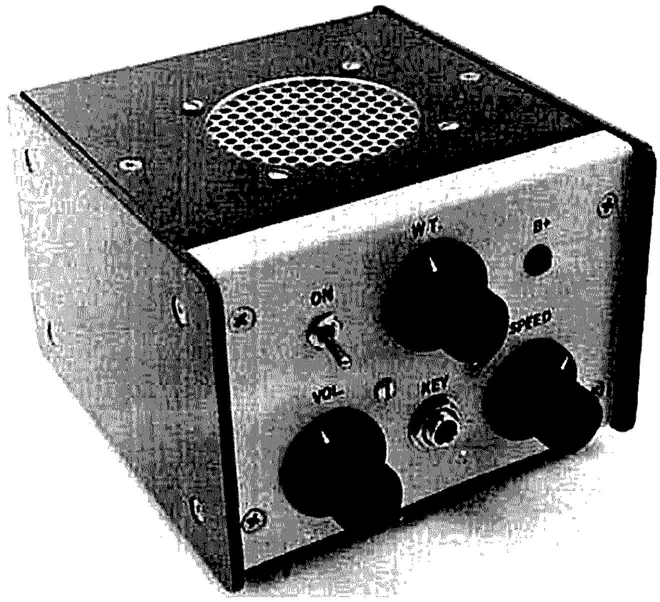
Reports Of "Tandem Satellite" QSOs using both 06 and 07 as their orbits coincided have been received by AMSAT, but more inputs are needed from those who have made contacts using the two transponders in tandem. Send QSO data to AMSAT, Box 27, Washington, D.C. 20044.

ANOTHER 10-METER BEACON will help 10-meter DXers through the predicted flat conditions ahead. The NZART has put ZL2MHF on 28170 kHz with 90 watts input and a half-wave vertical. ID is in CW every 10 seconds or so. Reports of reception are requested to NZART Upper Hutt Branch 63 Inc., Box 40212, Upper Hutt, New Zealand.

Proposed San Diego Beacon, widely reported elsewhere as on the air last fall, has still not been okayed by FCC. FCC request for further data back in October still has not been answered.

HR REPORT GOES WEEKLY. On March 1st, HR Report started going out weekly instead of twice monthly, more than doubling the number of issues per year to a minimum of 50. Change insures readers that "hot" news will be just that much hotter, gives us the opportunity to cover more stories and to cover important stories in greater depth.

Subscription Rates will remain unchanged, and only cutback will be from airmail to first-class mail for continental U.S. and Canadian subscribers.



integrated-circuit electronic keyer

This novel electronic keyer, based on the new Curtis 8043 keyer IC, features dot memory, self-completing dots, dashes and spaces, sidetone, and almost zero current drain as well as iambic keying

Hank Olson, W6GXN, Post Office Box 339, Menlo Park, California

Any who has built one of the many versions of IC keyers described in the amateur radio magazines probably has been less than satisfied with its performance in the shack. Many circuits use RTL as the basic logic form, but RTL has inherently poor noise immunity and requires large supply currents. Some newer designs use TTL – ICs which have improved noise immunity, but are still in the current-guzzler category. In virtually all the circuits that I have built or even thought about building (including a couple of my own design) there have been "glitches" in the designs. These have shown up in actual use by operators, and are subtle problems that can be difficult for a designer to anticipate.[1,2]

The Curtis 8043 is a new IC developed by an engineer who has been in the electronic keyer business for years, and who has also been employed by one of the largest IC manufacturers in the United States. Such a unique combination of experience should make the Curtis 8043 a nearly ideal IC around which to build your own keyer, and this has been proven by the keyer described here. First of all, the 8043 device uses

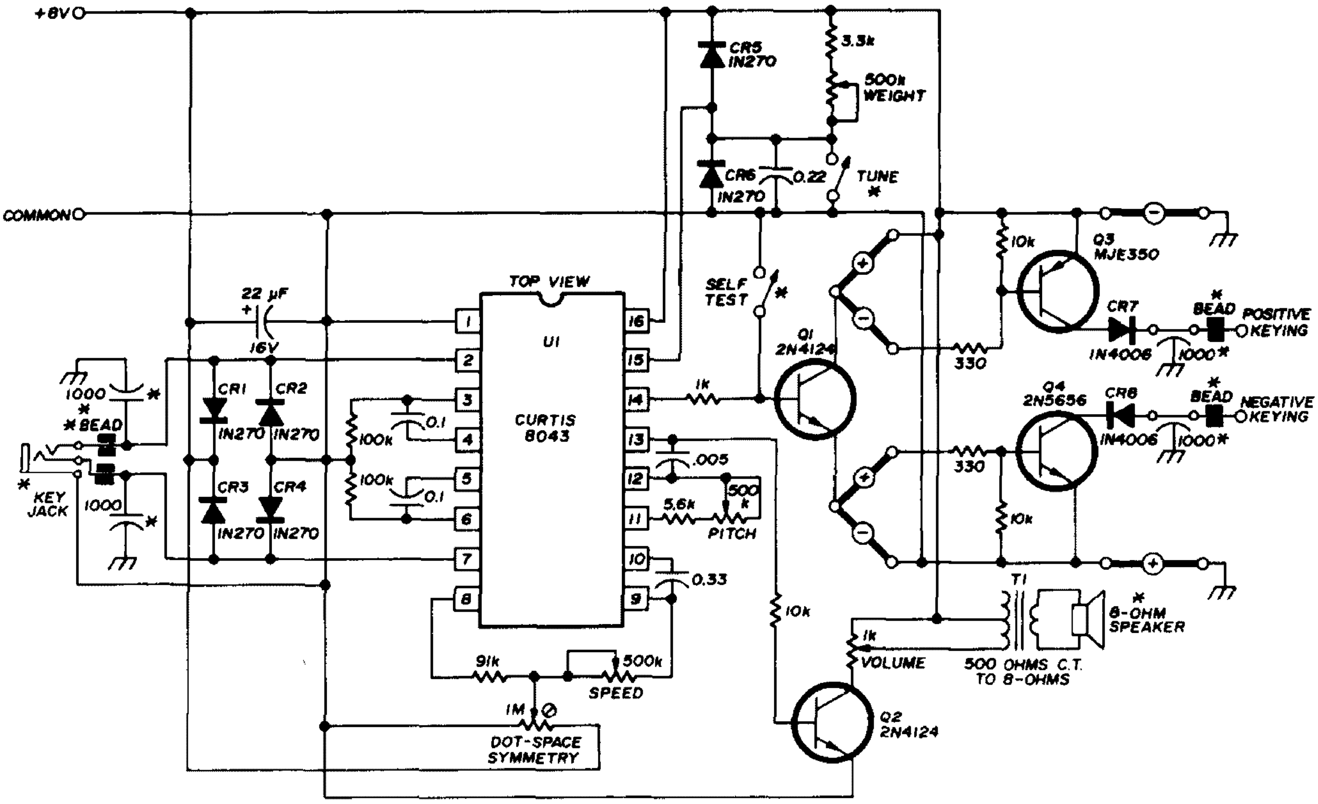


fig. 1. Circuit diagram of the electronic keyer using the Curtis 8043 IC. Items not provided in the 8043-1 partial kit are marked with an asterisk. Jumpers marked with plus sign are installed for positive key-up voltages; jumpers marked with negative sign are used for negative key-up voltages (see text). Circuit uses extensive decoupling to eliminate possible problems with RFI.

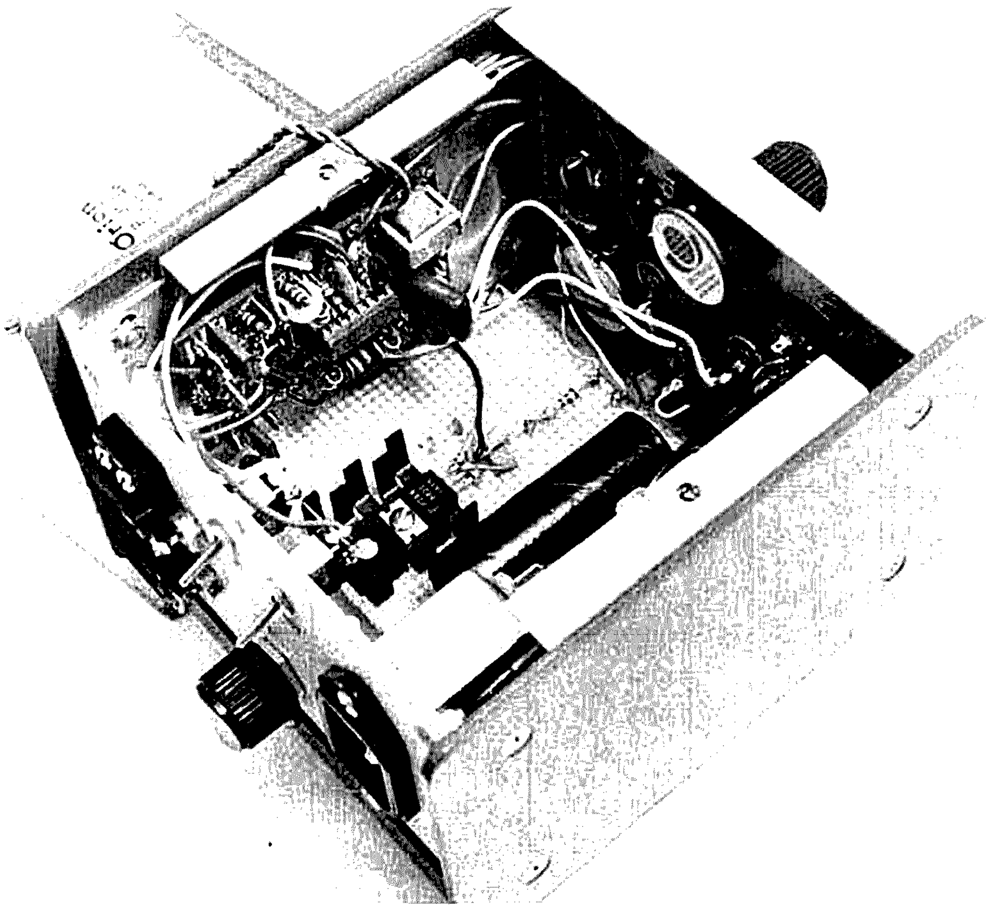
cmos logic, the ideal logic family for systems such as an electronic keyer. That is, for low speed logic (slower than 10 MHz) cmos provides low power consumption and high noise margin. As with most cmos ICs, a wide latitude for power-supply voltage is a feature; the Curtis 8043 will operate with supply voltages from +5 to +12 volts.

The whole gamut of CW operating features are included in the 8043 IC: self-completing dots, dashes and spaces, iambic keying option, dot memory, weight control, sidetone and almost zero current drain (50-100µA quiescent, 10-30 mA key down). In addition, solid-state keying is used in the recommended Curtis circuit, rather than relay keying, and provisions have been made for either positive or negative key-up voltages across the keying transistor (up to 300 volts). The recommended circuit (the one that comes with the 8043 partial kit*) was the one I used. The kit printed-circuit board is only about 1-1/2x2-3/4 inches (38x70mm) and its use with the recommended circuit saves a lot of construction time. The speaker, cabinet, jacks and power supply are *not* part of the partial kit and must be supplied by the builder. A simple power-supply circuit is presented in the manual that comes with the device, but I used a better-regulated supply of my own design (because I'm a regulated power-supply nut).

The keyer is built into a Moduline

*Partial kit 8043-2 includes the 8043 IC, PC board, IC socket, manual, 200 mA keying transistors, manual and most other small parts excluding speaker, chassis, jacks, etc. (asterisked items in fig. 1) and is priced at $49.95. Kit 8043-1, including the 8043 IC, PC board, IC socket and manual, is priced at $24.95. Order from Curtis Electro Devices, Inc., Box 4090 Mountain View, California 94040. See ad on page 82.

P355 cabinet with all input and output leads decoupled to prevent rf from getting into the circuitry. The 8043 is presumably comparatively rf-proof, but I have found that it's much easier to install RFI-proofing as I build, rather than to try to add it after the fact. The ac line is decoupled using a Corcom 6EF1 line filter; the keyer contacts and transmitter keying line are decoupled with 1000-pF feedthrough or standoff capacitors and ferrite beads. These RFI measures are not absolutely necessary, but probably contribute to the lack of trouble from rf that the unit enjoys.



Construction of the electronic keyer. The 8043 IC is installed on the small printed-circuit board on the side wall of the enclosure. The regulated power supply is built on the perf board at the bottom of the chassis.

circuit

The keyer circuit as furnished by Curtis is shown in fig. 1; asterisk-marked items are builder-supplied. The power-supply circuit is shown in fig. 2. Note that there are three jumper wires on the Curtis board; two are associated with Q1 and one with the emitters of Q3 and Q4, transistors used as switches to replace the relay commonly found in electronic keyers for keying the transmitter. The jumpers determine which transistor (Q3 or Q4) is used.

In this service keying transistors are better than relays because they have no contact bounce and no contacts to wear. However, the transistor switches do not have the complete dc isolation of a pair of relay contacts, so an npn device must be used for keying trans-

mitters having a positive key-up voltage (such as cathode keying) and a pnp device must be used for keying transmitters having a negative key-up voltage. Both Q3 and Q4 will stand up to 300 volts in the open-key condition.

As can be seen in the photographs of the IC keyer, the *speed, volume* and *weight* controls are mounted on the front panel with the ac switch, LED pilot light and key jack. The *pitch* control is mounted on the rear apron of the

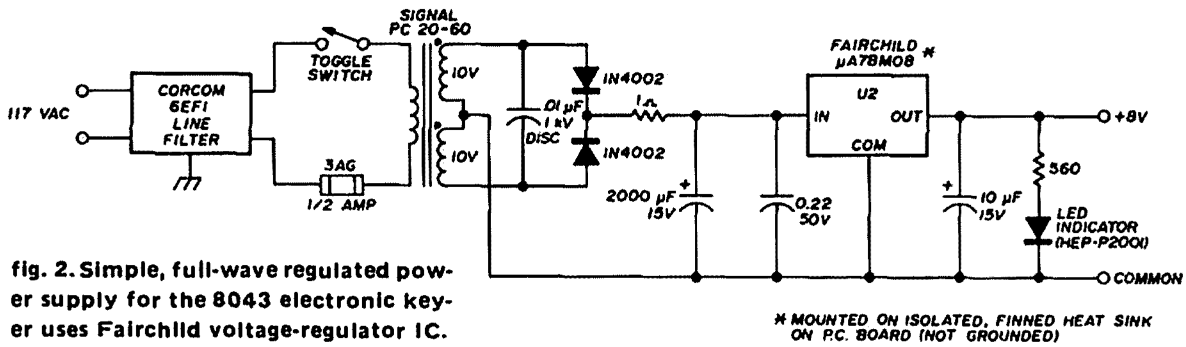
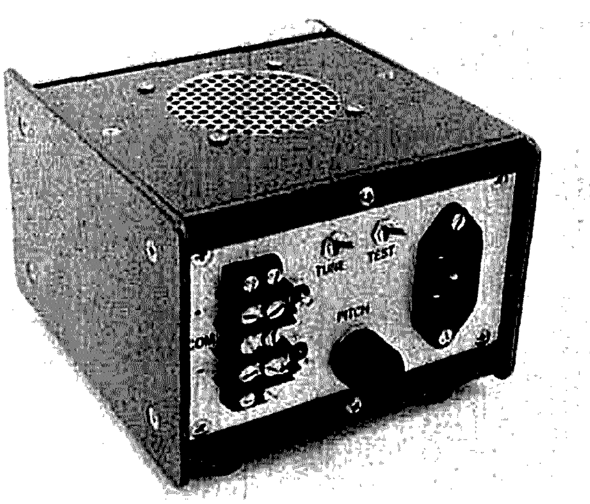


fig. 2. Simple, full-wave regulated power supply for the 8043 electronic keyer uses Fairchild voltage-regulator IC.

In addition to the line switch, there are two other miniature toggle switches, one labeled *tune*, the other *self-test*. The *tune* switch turns the transmitter on for tuning up, and the *self-test* feature allows you to operate the keyer (*listening* to its output) without keying the transmitter. If straight-key operation is desired, the key may be connected (with a jack) in parallel with the *tune* switch.

Rear view of the electronic keyer showing the tune and test switches, pitch control, and terminal strip for transmitter connections. Plug on the right is for the ac line. Monitor speaker is on top.



keyer along with the *tune* and *self-test* switches.

The 1-meg pot referred to in fig. 1 as *dot-space symmetry adjust* is a small board-mounted trimpot. This *symmetry* pot is the only adjustment necessary after the keyer has been built. Starting with the *symmetry* pot at the middle of its adjustment, and with the *weight* control at minimum and the keyer paddle on dots, the *symmetry* control is adjusted until the dot and space periods are equal, as displayed on an oscilloscope.

summary

The keyer built around the Curtis 8043 semi-kit has proven to be everything it claimed to be, satisfying all comers at a local amateur radio club. No signs of RFI have been evident at the club station nor at other 1000-watt amateur stations. If you've been thinking about building a keyer, the 8043 is clearly the way to go!

references

1. Howard Berlin, K3NEZ, "Memory Keyer," (letter), *ham radio*, December, 1974, page 58.
2. F.E. Hinkle, "Heath HW-7 Modifications," *ham radio*, December, 1974, page 60.

ham radio

microstripline preamplifiers for 1296 MHz

Complete design and construction information for low noise, solid-state preamplifiers for the amateur 23cm band

Paul Shuch, WA6UAM, 14908 Sandy Lane, San Jose, California 95124

In a previous article I described a low-cost transceive converter for the amateur 1296-MHz band.[1] Calculated parameters and preliminary measurements indicated a receive system noise figure of 7.5 dB. As is often the case, these original claims have proved overly optimistic. Subsequent measurements with an argon-discharge noise source and an automatic noise-figure meter showed the true ssb noise figure to be more on the order of 9.5 to 10 dB. Neither improvements in the homebrew balanced mixer nor the use of high-grade commercial mixers significantly altered this figure.

It became apparent that optimum performance would necessitate the use of one or more low-noise preamplifier stages preceding the diode balanced mixer. Several popular amplifier circuits[2,3,4] were tried and the results were entirely satisfactory. However, I wanted to depart from the conventional technology (pi-network input and output with slab inductors) used in these designs. Since I had obtained considerable success with microstriplines in a family of transmit linear amplifiers, I decided to build a family of receive preamps using the same basic design techniques. The resulting circuits, presented here, represent a reasonable tradeoff between cost, performance, simplicity and reproducibility.

system considerations

It was assumed that these amplifiers would precede a receive converter with a ssb noise figure on the order of 10 dB. The Simple Sideband System[1] meets this criteria, as do most properly adjusted trough-line converters,[5] and at least one popular commercial unit.*

A workable rule of thumb is the principle that, if the gain preceding a receive converter is at least 10 dB greater than the converter's noise figure, then the noise figure of the total system will, for all intents and purposes, equal the noise figure of the preamplifier. Thus, 20 dB of preamplifier gain preceding a 10-dB NF converter will have the effect of masking the converter's noise.

A number of readily available microwave transistors are capable of 2- to 3-dB noise figures in the 1296-MHz

*Spectrum International MM_c 1296, $85.95 from Spectrum International, Box 1084, Concord, Massachusetts 01742.

amateur band. These devices offer conservatively rated power gains on the order of 10 dB per stage so if two such stages of preamplification precede the 10-dB NF converter, an overall noise figure of 2 to 3 dB will result.

When receive preamplifiers are connected in cascade, the first stage should be adjusted for optimum noise figure; subsequent stages are adjusted for optimum power gain. In practice the only difference lies in the input matching circuit. When a power-match is desired, the input circuit presents a complex conjugate match to the transistor's input impedance. In the case of a noise-match, a predetermined mismatch is introduced into the input circuit to minimize the stage's noise figure. In both cases the transistor's collector should look into a complex conjugate match.

microstripline considerations

The exact details of a microstripline design vary with the material used for the substrate. Such dielectrics as Teflon, Rexolite, Duroid, and glass offer superior performance at microwave frequencies. The material most readily available to the experimenter, however, is fiberglass-epoxy printed-circuit board. After having designed and built more than a dozen preamplifiers of the type described here, on a variety of substrates, I can state conclusively that the degradation in performance resulting from building on lowly glass-epoxy board is beyond the measurement capabilities of the average experimenter. Thus the amplifiers presented here were designed to be etched onto 1/16-inch (1.5mm) thick G-10 PC board, double-clad with 1-ounce copper.

It is possible to design an amplifier using microstriplines that requires no external tuning — all the resistive and reactive matching elements are provided by the microstriplines. Such designs have been published previously, and if properly built, will provide excellent performance without the need for tuning adjustments. However, amplifiers without tuning adjustments are practically an affront to the amateur spirit. Though considered frivolous by some, my amplifiers include trimmer capacitors. In addition to giving the dyed-in-the-wool experimenter something to tweak, these adjustable components provide some degree of compensation for slight variations between transistors, as well as variations from one PC board to the next.

preamplifier transistors

The second-stage amplifier is built around a Hewlett-Packard 35826E, a low-cost version of the well known HP-21 family. An acceptable substitute is the VO21 manufactured by the Nippon Electric Company, a second-source device which performs identically in the circuit. At the time of this writing, both transistors sell for $17.50 each in single quantities.*

The device used in the optimum-noise-matched first stage depends on the needs and budget of the individual builder. The design presented here uses a Hewlett-Packard 35866E option 100, one of the least expensive members of the low-noise HP-22 family. Priced at $45, it provides an overall system noise figure of 2 dB, challenging the best parametric amplifiers of yesteryear. Unless you anticipate EME or long-haul troposcatter communications, you will probably find it more cost-effective to use the lower priced HP-21 or VO21 in

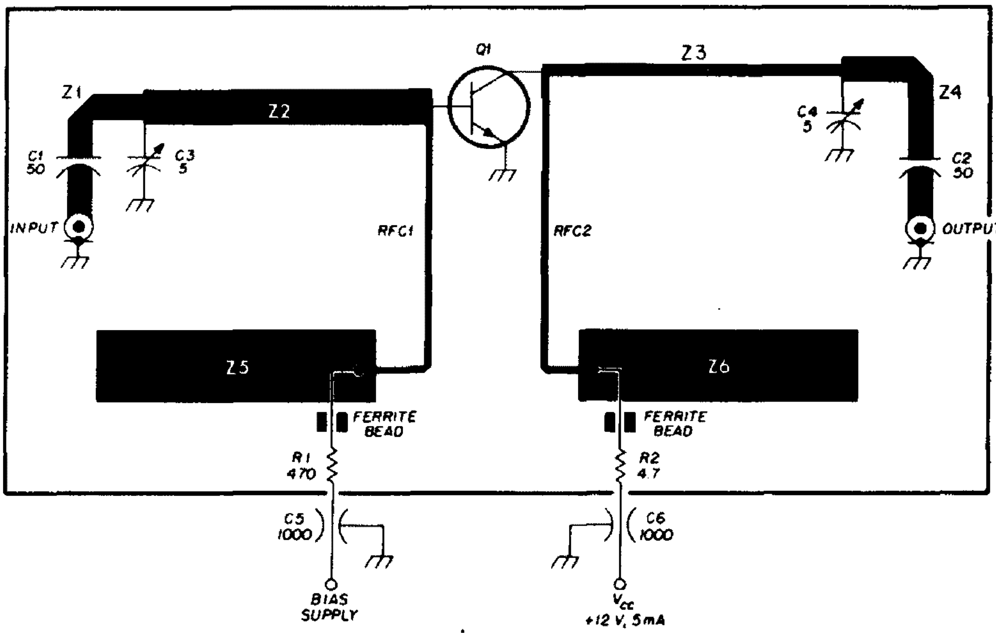
*For the address of your nearest Hewlett-Packard distributor, look in the *Yellow Pages* or write to *HPA Division*, 540 Page Mill Road, Palo Alto, California 94304. Nippon Electric Company semiconductors are distributed in this country by California Eastern Labs, Inc., 1 Edwards Court, Burlingame, California 94010.

the first stage as well. These devices, when tuned for optimum noise figure in the circuit shown in fig. 1, will yield an overall system noise figure on the order of 3 dB.

The cost factor can be easily analyzed in terms of dollar expenditure per dB improvement in signal-to-noise ratio. For example, two stages of HP-21 or VO21 at $17.50 per transistor will improve a 10-dB NF converter by 7 dB at a device cost of $5.00 per dB improvement. With the higher performance HP-22 in the front end, an additional dB of sensitivity can be achieved, but at an added cost of $27.50. The overall cost-effectiveness of the two-stage preamplifier now becomes $7.81 per dB — not an unreasonable figure when maximum performance is required.

Since it is common practice for microwave semiconductor manufacturers to produce a family of transistors by mounting the same chip in a variety of packages, it is often possible to substitute a different transistor of the same family without significantly influencing amplifier performance.

The part numbering system used at



C1,C2 50 pF chip capacitor (ATC 100 or equivalent

C3,C4 1 to 5 pF precision piston trimmer (Johannson JMC 4642)

C5,C6 1000 pF feedthrough capacitor

Q1 Hewlett-Packard 35866E, option 100 preferred (HP 35826E or NEC VO21 acceptable)

R1 470 ohm, ¼ watt carbon composition

R2 4.7 ohm, ¼ watt, carbon composition

RFC1 RFC2 100 ohm, quarter-wavelength microstripline, 0.02" (0.5mm) wide, 1.25 inch (32mm) long

Z1,Z4 50 ohm microstripline, 0.1" (2.5mm) wide, any convenient length

Z2 41.7 ohm, quarter-wavelength microstripline, 0.14" (3.5mm) wide, 1.18" (30mm) long

Z3 75 ohm, quarter-wavelength microstripline, 0.04" (1mm) wide, 1.22" (31mm) long

Z5,Z6 Rf short. 25 ohm, quarter-wavelength open-circuited microstripline, 0.30" (7.5mm) wide, (32mm) long

fig. 1. Circuit for the noise-matched 1296-MHz preamplifier stage. Printed-circuit layout is shown in fig. 2.

Hewlett-Packard provides some insight into the interchangeability of their devices. A system of five digits and a one-letter suffix is used, the first three digits being 358. The fourth digit position indicates the device family, with the number 6 designating HP-22 type devices (low noise microwave transistors), the number 2 referring to HP-21 type transistors (for general-purpose microwave applications), and the numbers 3 and 5 indicating the linear power transistors of the HP-11 and HP-12 families.

The fifth digit of the Hewlett-Packard part number indicates the type of package in which the semiconductor chip is mounted. Among the strip-packs, a number 2 indicates a 70-mil (1.8mm) diameter; 1, a 200-mil (5mm) diameter; and 6, a 100-mil (2.5mm) square package. The number 4 refers to a metal TO-72 package, and 7 is a coaxial package. Grounded-stud packages are designated by a 4, and 3 indicates a grounded-bar configuration. Unpackaged chips are coded zero. The letter suffix indicates whether the device is mounted in common-emitter (E) or common-base (B) configuration. Unpackaged chips carry the suffix A. The HP 35821E, for example, is a general-purpose (HP-21) microwave transistor mounted in a 200-mil (5mm) round strip package in the common-emitter configuration.

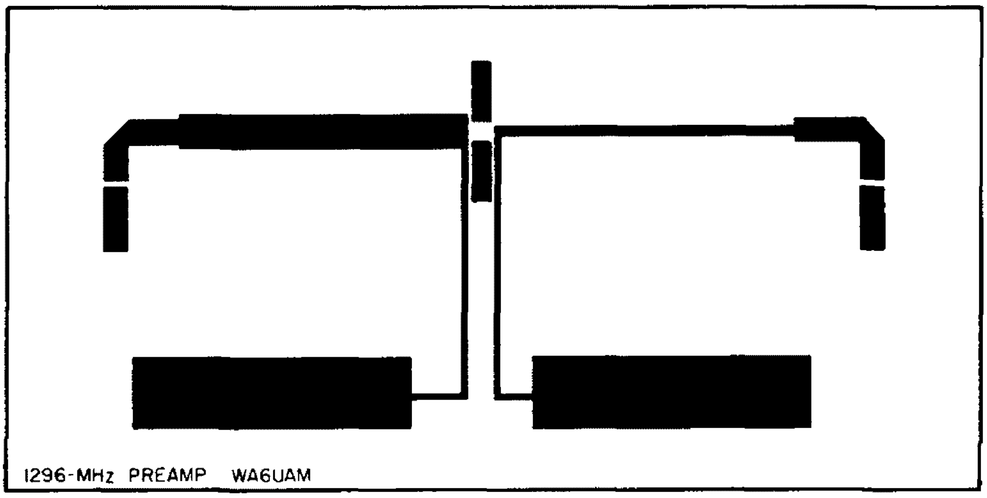


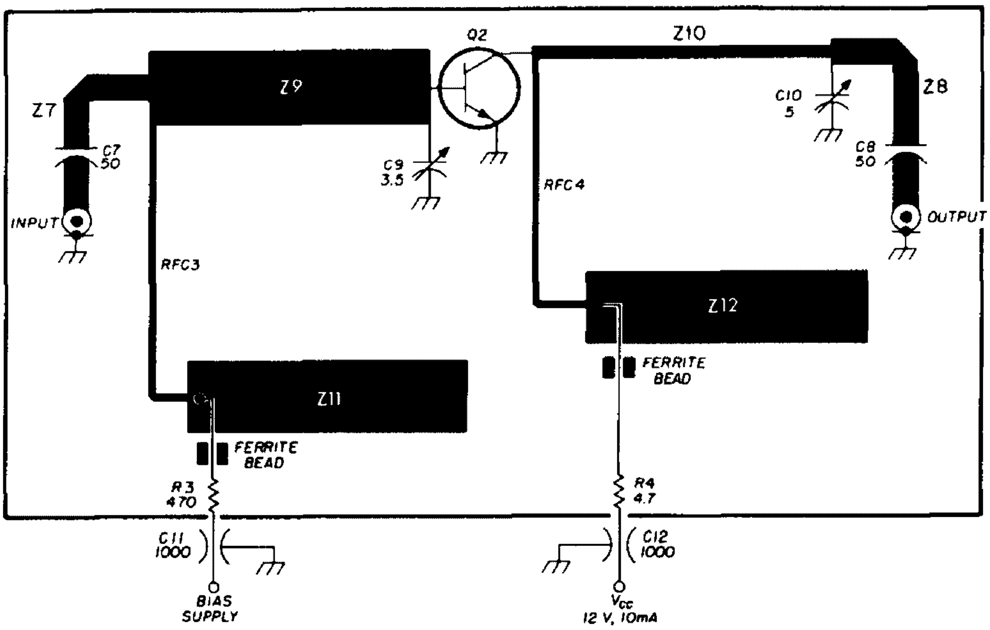
fig. 2. Full-size printed-circuit layout for the noise-matched 1296-MHz preamplifier shown in fig. 1.

Obviously, differently packaged versions of the same semiconductor chip will differ from one to another in terms of their complex input and output impedances. However, at low frequencies (and for microwave devices 1.3 GHz can be considered low), the contribution of package parasitics to the overall input (S_{11}) and output (S_{22}) S-parameter values can be considered small in relation to the characteristics of the chip itself.* These amplifiers, though designed around the characteristics of the 100-mil (2.5mm) square package, function well with like transistors in the 70-mil (1.8mm) and 200-mil

*The S-parameters or scattering parameters are one of several methods used for describing the input-output characteristics of transistors. The basic difference between S-parameters and other systems are that the S-parameters are measured with transmission-line inputs and outputs; this greatly simplifies measurements at uhf and microwave.

(5mm) round packages with only minor degradation in input and output vswr.

An analogous situation exists with Nippon Electric Company transistors, although NEC lacks the elaborate numbering system which identifies the HP with this newer transistor in the existing boards, performance was wholly satisfactory. As a matter of fact, the VO21 in the 100-mil (2.5mm) square package exhibited 0.2 dB lower noise figure than the original device.



C7,C8	50 pF chip capacitor (ATC 100 or equivalent)
C9	0.35 to 3.5 pF precision piston trimmer (Johannson JMC 5801)
C10	1 to 5 pF precision piston trimmer (Johannson JMC 4642)
C11, C12	1000 pF feedthrough capacitor
Q2	Hewlett-Packard 35826E or NEC VO21
R3	470 ohm, ¼ watt carbon composition
R4	4.7 ohm, ¼ watt carbon composition
RFC3 RFC4	100 ohm, quarter-wavelength microstripline, 0.02" (0.5mm) wide, 1.25" (32mm) long
Z7,Z8	50 ohm microstripline, 0.10" (2.5mm) wide, any convenient length
Z9	25 ohm, quarter-wavelength microstripline, 0.30" (7.5mm) wide, 1.14" (29mm) long
Z10	75 ohm, quarter-wavelength microstripline, 0.04" (1mm) wide, 1.23" (31mm) long
Z11, Z12	Rf short. 25 ohm, quarter-wavelength open-circuited microstripline, 0.30" (7.5mm) wide, 1.14" (29mm) long

fig. 3. Circuit for gain-matched 1296-MHz amplifier stage. Printed-circuit layout for this amplifier is shown in fig. 4.

devices. The NEC VO21 was initially manufactured in a 150-mil (3.8mm) round package, and it was the scattering parameters for that device which were used in the design of these amplifier boards. Recently NEC introduced a VO21 chip mounted in a 100-mil (2.5mm) square package. Although perfect matching could not be achieved

When ordering the NEC VO21 transistor, specify the 320 package if a 150-mil (3.8mm) round package is desired. For the 100-mil (2.5mm) square, request package ML-3. There is no price difference.

construction

Figs. 1 and 3 are functional sche-

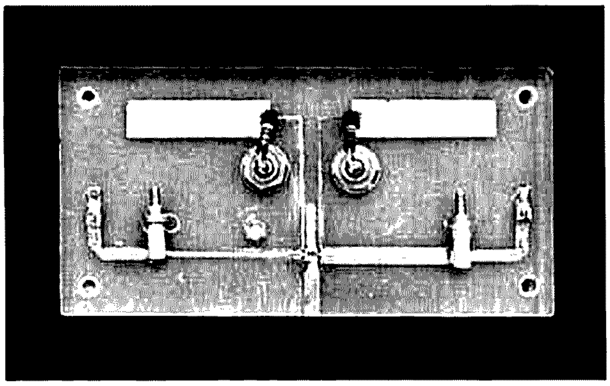
matics of the amplifier stages. Microstriplines opposite a groundplane (the unetched side of the double-clad printed-circuit board) comprise all matching transformers, rf chokes and rf bypasses. **Figs. 2** and **4** are full-size printed-circuit layouts for the noise-matched and power-matched stages, respectively. Note that dimensions are applicable only to 1/16-inch (1.5mm) G-10 double-sided glass-epoxy printed-circuit board.

Mounting of the transistors is detailed in fig. 5. A hole is drilled in the PC board to allow direct strapping of the emitter traces to the groundplane. It is desirable to cover this hole on the groundplane side after the transistor is installed, to improve shielding and furnish physical protection for the transistor.

A word of caution is in order regarding mounting the transistors to the boards. There is a difference in lead layout between the HP and NEC transistors which can cause considerable confusion. The emitter leads of both devices are the wider pair of leads. One of the two remaining leads is tapered 45° at the end. The slashed lead is the *base* of the NEC devices, but the *collector* of the HP transistors. This small difference cost me two expensive transistor failures while developing the prototype amplifiers for this article.

The grounded (adjusting screw) end of the trimmer capacitors must be connected directly through the board to the groundplane to minimize tuning-tool interactions when making adjustments. Although concentric-ring piston trimmers (Johannson, JFD, etcetera) are ideal for this purpose, performance of the lower cost ceramic piston trimmers used in uhf TV tuners is entirely satisfactory.

The dc blocks at the input (C1 and C7) and output (C2 and C8) should ideally be ceramic chip capacitors. If these are not available, modified miniature disc ceramic capacitors are usable. An Xacto knife is used to scrape the insulation off of the sides of the capacitor, the leads are removed and, using the lowest possible soldering heat, the PC traces are bridge-soldered directly to the capacitor plates.



Microstripline side of the noise-matched 1296-MHz preamplifier. The low-noise H-P 35866E transistor is at the top center of the board. The gain-matched stage is similar except for the width of the input quarter-wave transformer and placement of the input trimmer capacitor.

The prototype preamplifiers shown in the photographs use miniature SMA coaxial connectors, but the less expensive JCM connectors made by E. F. Johnson are satisfactory substitutes. TNC connectors may be used, too, but they are much larger. BNC connectors should be avoided as they will slightly degrade the noise-figure performance of the amplifier stages. **Fig. 6** shows a method of modifying flange-type bulkhead connectors for use as microstripline launchers.

All power-supply leads for the collector supply and bias current are isolated from the rf circuitry by rf chokes, ferrite beads and bypass and feedthrough capacitors. Therefore, the necessary bias circuitry can be installed on the groundplane side of the circuit board.

bias circuits

With the transistors specified, optimum noise figure occurs at a collector current of approximately 5 mA. A higher collector current will improve stage gain. In the conventional configuration of a low-noise first stage and higher-gain second stage, it is desirable to bias the two stages for collector currents of 5 and 10 mA, respectively.

Virtually any bias circuit which will maintain the desired collector current is acceptable. Many of the simpler resistive bias circuits should be avoided due to their low stability factor (that is, high dependence of collector current on transistor dc current gain) and the resulting danger of thermal runaway. To quote a useful Hewlett-Packard applications note, "Often the least considered factor in microwave transistor circuit design is the bias network. Considerable effort is spent in measuring S-parameters, calculating gain, and optimizing bandwidth and noise figure, while the same resistor topology is used to bias the transistor. Since the cost per dB of microwave gain or noise figure is so high, the circuit designer cannot afford to sacrifice rf performance by inattention to dc bias considerations."[6]

An active bias circuit (variable constant-current source) is desirable in that it affords a degree of protection for the transistor while permitting ready collector current adjustment for optimizing stage gain and/or noise figure. One such circuit, shown in fig. 7, furnishes a variable collector current of 2 to 12 mA, more or less independent of the dc current gain of the transistor being biased.

For initial amplifier tuneup, adjust the trimpot in the base of the bias transistor to produce 5.5 mA of total current in the first amplifier stage, and 10.5 mA in the second. (This accounts for a quiescent bias circuit current of 0.5 mA). Upon completion of all rf tuning, the trimpots may be adjusted to optimize the overall system noise figure.

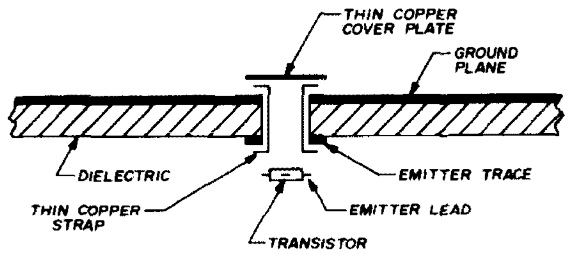


fig. 5. Method for mounting the transistors to the printed-circuit boards. Note thin copper strap used to connect the emitter circuit trace to the ground plane. Solder emitter leads as close as possible to the transistor package and use minimum soldering heat to avoid transistor damage.

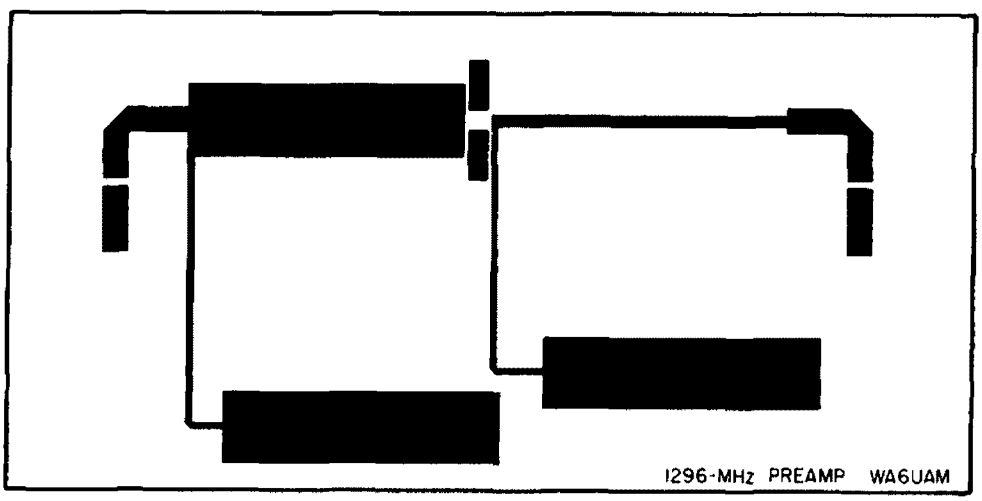


fig. 4. Full-size printed-circuit layout for the 1296-MHz gain-matched amplifier shown in fig. 3.

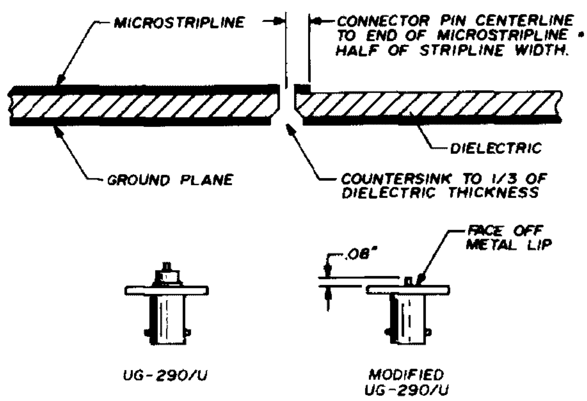


fig. 6. Modifying coaxial bulkhead connectors for use as microstripline launchers.

alignment

The two preamplifier stages are built as separate subassemblies and connected together with a short length of coaxial cable. The reasons for this modular approach are twofold. Most obvious is the fact that many operators may wish to add a single stage of preamplification initially, expanding their system later as needs and budget dictate. In this case it is recommended that the power-gain matched stage (which uses the less expensive transistor) be built first.

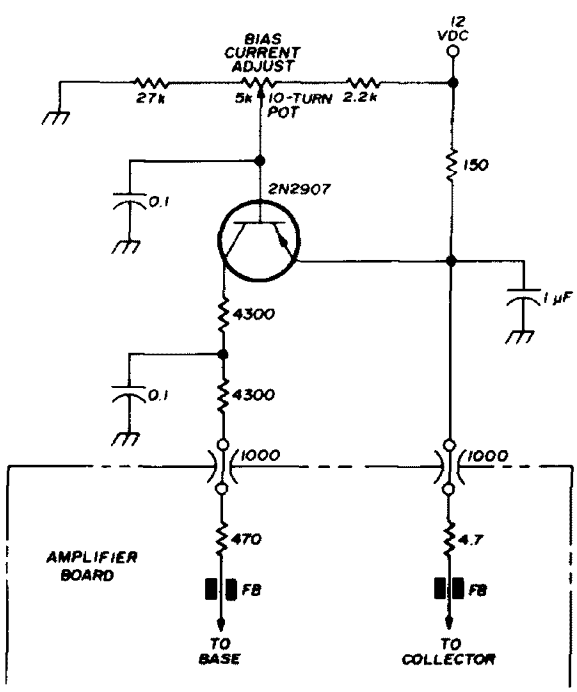


fig. 7. Active bias circuit provides adjustable collector currents from 2 to 12 mA. Components below dotted line are part of the amplifier circuit (see figs. 1 and 3).

Even if two stages of preamplification are built, tuneup and matching can be most readily accomplished if the stages are in separate modules. The power-matched stage is connected to the receiving converter first, its input terminated in 50 ohms, and the trimmer capacitors adjusted for maximum received signal from a beacon source.[7,8,9] Once a power match is achieved in the second stage (both input and output), the first stage is connected and, using a weak-signal source, its *output circuit only* is adjusted for maximum received signal. A power match now exists between the two amplifiers as the output of the first stage and the input to the second stage are each matched to 50 ohms, and a 50-ohm coaxial cable connects the two.

The tuning process is completed by obtaining a proper noise match into the first stage. This is most readily accomplished by using an argon-discharge type noise source.[10,11] Unfortunately, few experimenters have access to such equipment, except perhaps at regional uhf conferences.* A semiconductor diode noise source is an acceptable alternative.[12,13] A number of articles have described the process of tuning an amplifier for minimum noise figure.[14-18]

When adjusting the input circuit of the first stage, an important consideration is the interactive nature of the input matching and bias current adjustments. Since the transistor's optimum source reflection coefficient, Γ_o (see page 24), varies with collector current,

*One of the more popular attractions at the annual West Coast VHF Conference is a receiver noise-figure competition, during which participants may optimize the performance of their receiving equipment with an automatic noise measuring system. A gas-discharge noise source is usually available for noise-figure measurements on both 1296 and 2304 MHz.

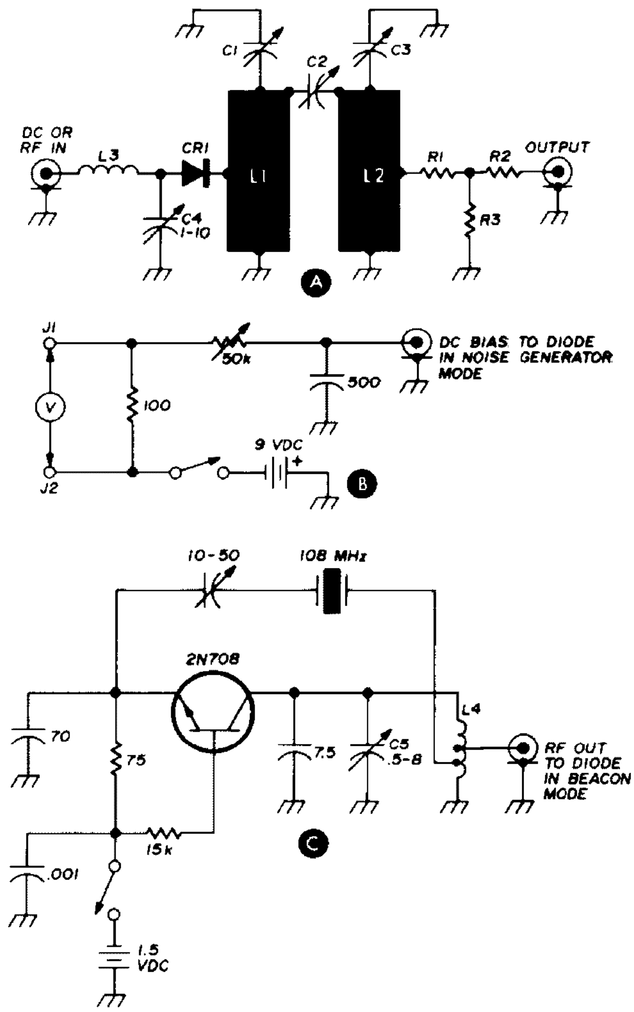
the ultimate in performance is achieved by alternate adjustments to first the input trimmer capacitor, then the bias pot. These adjustments are repeated until no further improvement in noise figure can be achieved.

The tuning elements of these amplifiers are inherently broadband. Without adequate selectivity at the input, bipolar transistors can exhibit disastrous overload characteristics. The cross-modulation and intermodulation effects of operating these preamplifiers in a high rf-density environment (i.e., virtually any populated area of the world today) can completely nullify any system noise-figure improvement. As a precaution it is good practice to provide input selectivity external to the preamplifier in the form of a high-Q filter in series with the antenna input. Single-pole coaxial or trough-line resonators, as well as some multi-pole microstripline filters, can provide adequate selectivity against out-of-band signals with a minimum of insertion loss (which would add to the amplifiers's noise figure in establishing receiver sensitivity).

When an input filter is used it is important to first optimize the preamplifier's noise figure as discussed above, then adjust the filter for minimum insertion loss at the operating frequency in a 50-ohm system. After mating the two, no further adjustments should be made to either the filter or the preamplifier unless a precision automatic noise-figure meter is available.

multipurpose uhf tuning instrument

It is interesting to note the similarities between the diode noise sources and rf beacons commonly used for adjusting uhf receiving preamplifiers. Both instruments normally have an output circuit consisting of a microwave diode feeding a tuned circuit, with an output port matched to 50 ohms. The two pieces of test equipment differ in that the diode is fed with direct current



C1,C2	0.5 to 2 pF piston trimmer capacitor
C3	(low-cost ceramic type acceptable)
CR1	microwave diode (1N25 or equivalent)
L1,L2	grounded microstripline, 0.3" (7.5mm) wide, 0.9" (23mm) long, tapped 0.2" (5mm) up from grounded end
L3	6 turns no. 20, 0.1" (2.5mm) diameter, 0.5" (13mm) long
L4	5 turns no. 20, 0.25" (6.5mm) diameter, 0.5" (13mm) long, tapped at 1 and 2½ turns from grounded end
R1,R2	10 ohm, ¼ watt carbon composition
R3	110 ohm, ¼ watt carbon composition

fig. 8. Combination weak-signal source (C) and diode noise generator (B). Microstriplines L1 and L2 are etched on 1/16" (1.5mm) double-clad, fiberglass-epoxy circuit board (A).

when used as a noise generator, versus rf when used as a weak signal source. The diode functions as a white-noise generator in the former application as opposed to harmonic generation in the latter.

There is no reason why the two

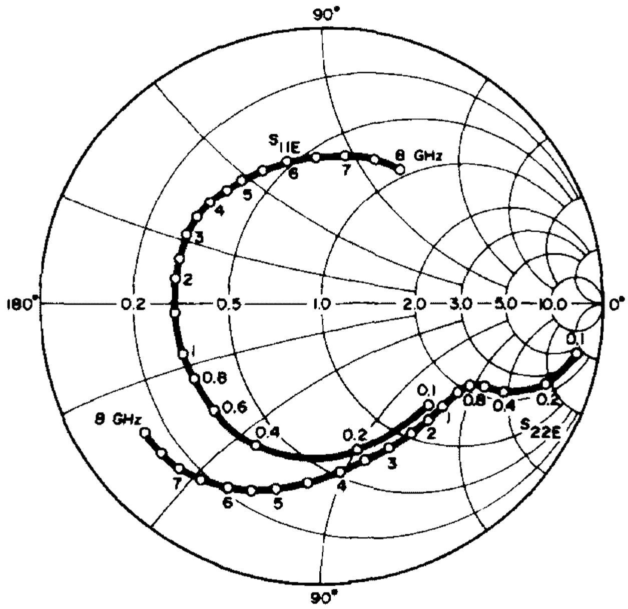


fig. 9. Input reflection coefficient, S_{11}, and output reflection coefficient, S_{22}, vs frequency for the Hewlett-Packard 35866E transistor (V_{CE} = 10 volts, I_C = 10 mA).

functions cannot be combined into a single instrument. **Fig. 8** represents an attempt to do so. The circuitry associated with the diode and its two-pole resonator is similar to the scheme I used in the local-oscillator section of the 1296-MHz transceive converter with filtering consisting of microstripline inductors etched on G-10 glass epoxy PC board. The three resistors at the output port form a 3-dB T-attenuator which assures a reasonable approximation of a 50-ohm output impedance. (In the weak-signal-source mode, additional pads may be placed between the output of the generator and the input to the receiver.)

The diode's input circuit is an adaptation of the familiar L-match used in the multiplier circuits of trough-line converters. The crystal-controlled 108-MHz oscillator which feeds the diode in the signal-generator mode is borrowed from a popular signal source design[7] while an existing noise-generator circuit[17] provided a workable diode biasing scheme. In short, the design presented in **fig. 8** is an amalgamation of various pieces of uhf test equipment into a single package.

Tuneup of the multipurpose instrument consists of merely connecting it in the rf beacon mode and tuning capacitors C1, C2, C3, C4 and C5 for maximum 1296-MHz signal into your receiver.

circuit design

For those readers who care to follow the calculations involved, the remainder of this article documents the procedure used to design the microstripline matching circuits of these 1296-MHz amplifiers. It should be pointed out that there are at least as many different methods for designing microstripline amplifiers as there are microwave engineers, and no one technique is necessarily any better or more workable than the others. The method shown here represents nothing more sacred than my own personal preference. It should be pointed out, however, that many of the more elegant amplifier designs used in the microwave industry are so complex as to be solvable only with the aid of a large digital computer. The designs shown here, though somewhat less precise, can be calculated by the average experimenter using only a slide rule.

Let us first consider a method for obtaining a complex conjugate impedance match to the amplifier transistors. Toward the end of this article I will discuss the special case of precisely mismatching the input to the first stage to obtain optimum noise figure.

Nearly all microwave semiconductor manufacturers publish Smith charts depicting the complex input and output impedances (S_{11} and S_{22}, respectively) as a function of frequency. **Figs. 9, 10** and **11** show such data for the HP 35866E, HP 35826E and NEC V021, respectively. In addition to these charts, most manufacturers furnish tabulated data listing the input and output impedances at various frequencies and differ-

ing bias conditions. It is important to note that these impedances vary significantly with changes in the dc operation of the transistor.

When tabular data is furnished, complex impedances are generally shown in polar form, i.e., magnitude and angle. This polar form may be converted to the more familiar rectangular notation ($A \pm jB$) on a Smith chart as indicated in **fig. 12**. Note that the magnitude is a decimal indication of the distance along a radius of the Smith chart, from zero (center) to 1 (circumference). The angle listed represents the direction of that radius. (For further material on the use of Smith charts see reference 19.)

Table 1 lists complex impedances, in both polar and rectangular form, for the transistors used in these amplifiers under the applicable dc bias conditions, at 1296 MHz. It is useful to convert the complex series impedances to their shunt equivalent circuit values. The applicable formulas are

$$R_p = R_s + \frac{X_s^2}{R_s}$$

$$X_p = \frac{R_s \times R_p}{X_s}$$

where R_s and X_s represent the resistive and reactive components, respectively, of the complex series impedance, and R_p and X_p represent the components of the parallel equivalent circuit. The parallel equivalents are included in table 1.

table 1. Typical S-parameters at 1.3 GHz (given in polar form, rectangular form and as parallel shunt equivalent, respectively).

	HP 35826E $I_c = 5$ mA	HP 35826E $I_c = 10$ mA	HP 35866E $I_c = 5$ mA	NEC V021 $I_c = 10$ mA
Input reflection coefficient, S_{11}	—	$0.61 \angle 178°$ $12.5 + j0.5$ $12.77 \parallel +j325.6$	—	$0.626 \angle 171°$ $11.75 + j4.0$ $13.11 \parallel -j38.5$
Output reflection coefficient, S_{22}	$0.57 \angle -41°$ $75 - j80.0$ $160 \parallel -j150$	$-0.51 \angle -40°$ $76 - j67.5$ $136 \parallel -j153$	$0.61 \angle -37°$ $80 - j90.0$ $181 \parallel -j161$	$0.266 \angle -65°$ $55 - j27.5$ $69 \parallel -j138$

output matching

It was desired to develop a single output circuit which would approximate a match to 50 ohms for any of the three referenced transistors, under either of the two indicated bias conditions. Assume for a moment that the reactive component of each transistor's parallel-equivalent complex output impedance is cancelled by a reactance of like magnitude and opposite sign shunting the collector (this will be accomplished shortly). Under these conditions the impedance to be matched to 50 ohms is a real value of magnitude R_p. For the conditions being considered, the value of R_p varies between 68.75 and 181.0 ohms. A compromise output circuit should match to the geometric mean of these outside values

$$R_p\ (\text{mean}) = \sqrt{R_p\ (\text{max}) \times R_p\ (\text{min})} = 111.6 \text{ ohms}$$

A 111.6-ohm nonreactive source may be matched to a 50-ohm nonreactive load through a quarter-wave matching transformer of characteristic impedance

$$Z_o = \sqrt{Z_{in} \times Z_{out}} = \sqrt{111.6 \times 50} = 74.7 \text{ ohms}$$

Still assuming no reactive component, the actual amplifier output impedance resulting from the use of each transistor

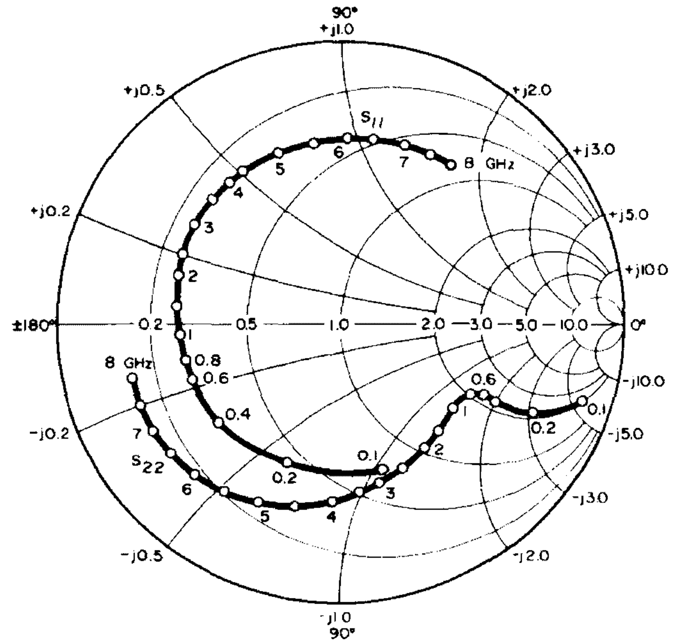


fig. 10. Typical input reflection coefficient, S_{11}, and output reflection coefficient, S_{22}, vs frequency for the H-P 35826E transistor (V_{CE} = 15 volts, I_C = 15 mA).

in this circuit would be that transistor's equivalent parallel output resistance transformed through a 75-ohm quarter-wave section. These values, along with

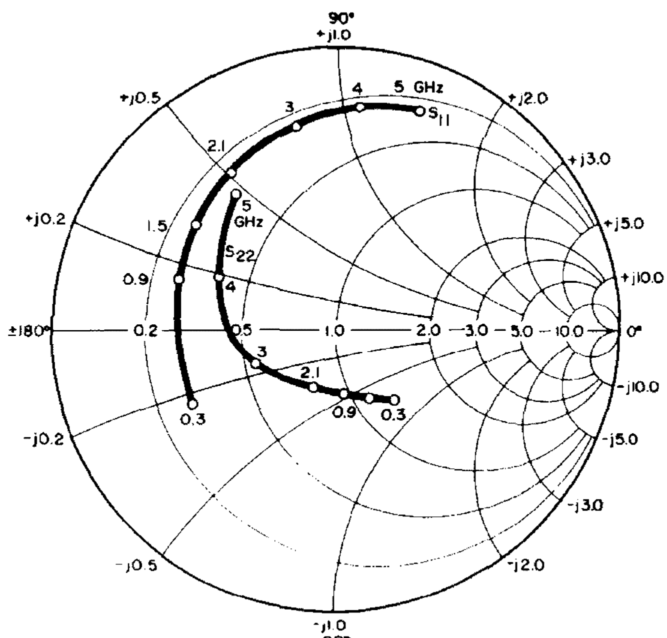


fig. 11. Typical input reflection coefficient, S_{11}, and output reflection coefficient, S_{22}, vs frequency for the NEC VO21 transistor (V_{CE} = 10 volts, I_C = 10 mA).

the resulting output vswr, are listed in table 2. Note that, even though restricted to a single output circuit for various combinations of device and collector current, this results in an acceptably low vswr well below 2:1.

Now, what about the reactive component of the transistor's output impedance so blithely ignored up to this point? Table 1 reveals the various values of parallel equivalent reactance, X_p, to be a shunt capacitive reactance (−j) in all cases, varying between 137.5 and 161.0 ohms. Obviously these capacitive reactances could be cancelled out by a variable inductor of like reactance range connected in shunt with the transistor's collector. However, from a practical standpoint, a *variable capacitor is a more* desirable tuning element than a variable inductor.

table 2. Actual output impedance and vswr of the Hewlett-Packard and NEC transistors in the circuits of figs. 1 and 3.

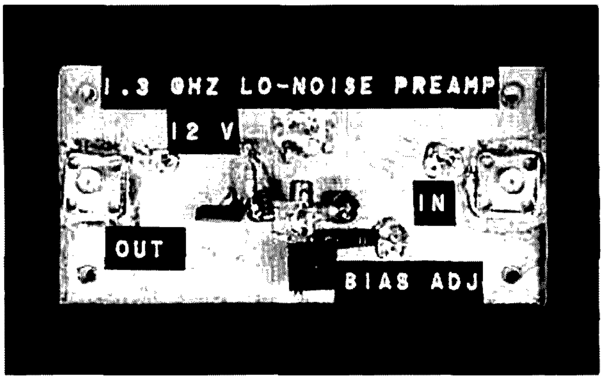
device	I_c	$Z_{out} = 75^2/R_p$	vswr
HP 35826E	5 mA	35.16 ohms	1.42:1
HP 35826E	10 mA	41.36 ohms	1.21:1
HP 35866E	5 mA	31.08 ohms	1.61:1
NEC V021	10 mA	81.82 ohms	1.64:1

The desired inductance may be realized by connecting a shunt capacitance to the collector through a quarter-wave transformer. As luck would have it, a quarter-wave transformer already exists at the collector circuit — the 75-ohm section used to match the transistor's parallel equivalent resistance to 50 ohms!

What value of capacitive reactance must be connected to the load end of the 75-ohm quarter-wave transformer? It must match the inductive reactance, X_{out}, resulting from transforming the transistor's shunt capacitive component, X_p, through a 75-ohm quarter-wave section. The relationship is

$$X_{out} = -\frac{Z_o^2}{X_{in}}$$

which results in a range of values for X_{out} of between +j34.9 and +j40.9 ohms. The required values for the trimmer capacitor, from $C = 1/2\pi f X_c$, is 3 to 3.5 pF at 1296 MHz. Allowing for variations between devices and imperfections in the micro-striplines, a 1 to 5 pF trimmer will assure cancellation of any of the transistors' shunt reactive components. The resulting compromise output circuit is used in the circuits of figs. 1 and 3.



Ground-plane side of the preamplifier stage. Note that component placement in the dc bias circuitry is not critical but the feed-through capacitors make convenient stand-offs. The transistor mounting hole is covered with copper foil to promote good rf shielding.

input matching

This discussion applies only to the second stage as a complex conjugate match into the first stage is not desired. From table 1, the input shunt reactance values for the two different transistors are found to be +j38.5 and +j325.6 ohms. As these values are positive (inductive), a shunt variable capacitor directly at the base will provide reactive matching. The capacitance range required is 0.38 to 3.2 pF. To provide sufficient tuning range, a 0.35 to 3.5 pF capacitor was chosen.

With the shunt reactive component thus disposed of, input matching now amounts to transforming 50 ohms to the required equivalent parallel resistance, R_p. The geometric mean of the R_p values for the two transistors is

$$R_p \text{ (mean)} = \sqrt{12.77 \times 13.11} = 12.94 \text{ ohms}$$

The required quarter-wave matching section has a desired characteristic impedance of

$$Z_o = \sqrt{12.94 \times 50} = 25.4 \text{ ohms}$$

Fig. 3 reflects these values. The resulting amplifier input impedance, as shown in table 3, yields an input vswr of better than 1.1:1.

noise-matching the first stage input

A parameter frequently specified for low-noise microwave transistors is the source reflection coefficient for optimum noise figure, Γ_s opt, or simply Γ_o, optimum source reflection coefficient. Either designation is an indication of the *source* impedance seen by the transistor input terminals when matched for optimum noise figure. The reference plane is at the edge of the transistor package, perpendicular to the input lead. The optimum source reflection coefficient, Γ_o, is often plotted as a function of frequency on a Smith chart, as indicated in fig. 13. It is important to realize that the impedances read from the Γ_o Smith chart are those looking back toward the source, *not* the transistor's input impedance.

If the input impedance to the amplifier (typically 50 ohms non-reactive) can be transformed to appear at the transistor input as Γ_o, optimum noise

table 3. Actual input impedance and input vswr of Hewlett-Packard and NEC transistors in the circuit of fig. 3.

device	$Z_{in} = 25^2/R_p$	vswr
HP 35826E	48.9 ohms	1.02:1
NEC V021	47.7 ohms	1.05:1

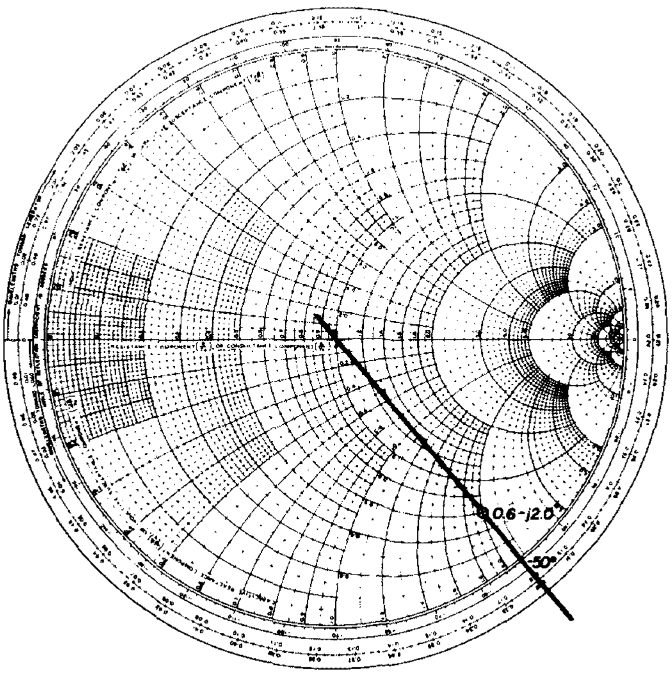


fig. 12. Using the Smith chart to convert impedance in the polar form to rectangular form (A ±jB). The magnitude and angle 0.8 $\angle -50^\circ$ lies on a radius passing through the -50° point on the outer circumference, 80% of the linear distance from the center to the edge of the chart, at 0.6 − j2.0 on the normalized Smith chart shown here (30 −j100 ohms in a 50-ohm system).

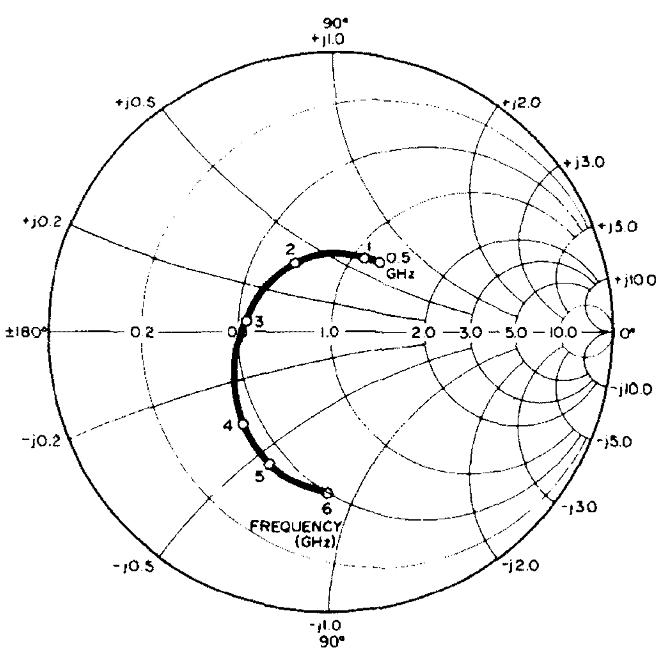


fig. 13. Typical optimum source reflection coefficient, Γ_o vs frequency for the Hewlett-Packard 35866E transistor (V_{CE} = 10 volts, I_C = 5 mA).

figure will result. **Fig. 13** shows that, for an HP 35866E option 100 device operated with I_C = 5 mA, Γ_o at 1.3 GHz is equal to 40 + j25 ohms. This translates to a shunt equivalent circuit of 34.8 ohms in parallel with + j55.6 ohms. Disregarding the inductive reactance shunting the base for a moment, the required real component can be readily realized by transforming the 50-ohm input impedance through a quarter-wave section with a characteristic impedance

$$Zo = \sqrt{50 \times 34.8} = 41.7 \text{ ohms}$$

As before, the desired inductive reactance shunting the transistor is achieved by adding a shunt capacitive reactance a quarter wavelength away from the transistor. The required capacitive reactance is

$$X_C = \frac{Z_o^2}{X\Gamma_o} = 31.3 \text{ ohms}$$

which at 1.3 GHz would represent a capacitance of 3.9 pF. A 1 to 5 pF trimmer is used for tuning the input to the first stage for optimum noise figure as shown in **fig. 1**.

modified interstage circuit

A number of the active 1296 operators who reviewed preliminary copies of this manuscript expressed an interest in a modified interstage design which would allow both stages of preamplification to be combined on a single amplifier board. Although I personally prefer separate modules, I concede that such a design would be of some value. A two-stage 1296-MHz preamplifier design by W6KQG[20] transformed the output impedance of one HP-21 into the complex conjugate of the second HP-21's input impedance through a single quarter-wave microstripline. A similar approach for these amplifiers, with the added provision of reactive tuning, is shown in **fig. 14**.

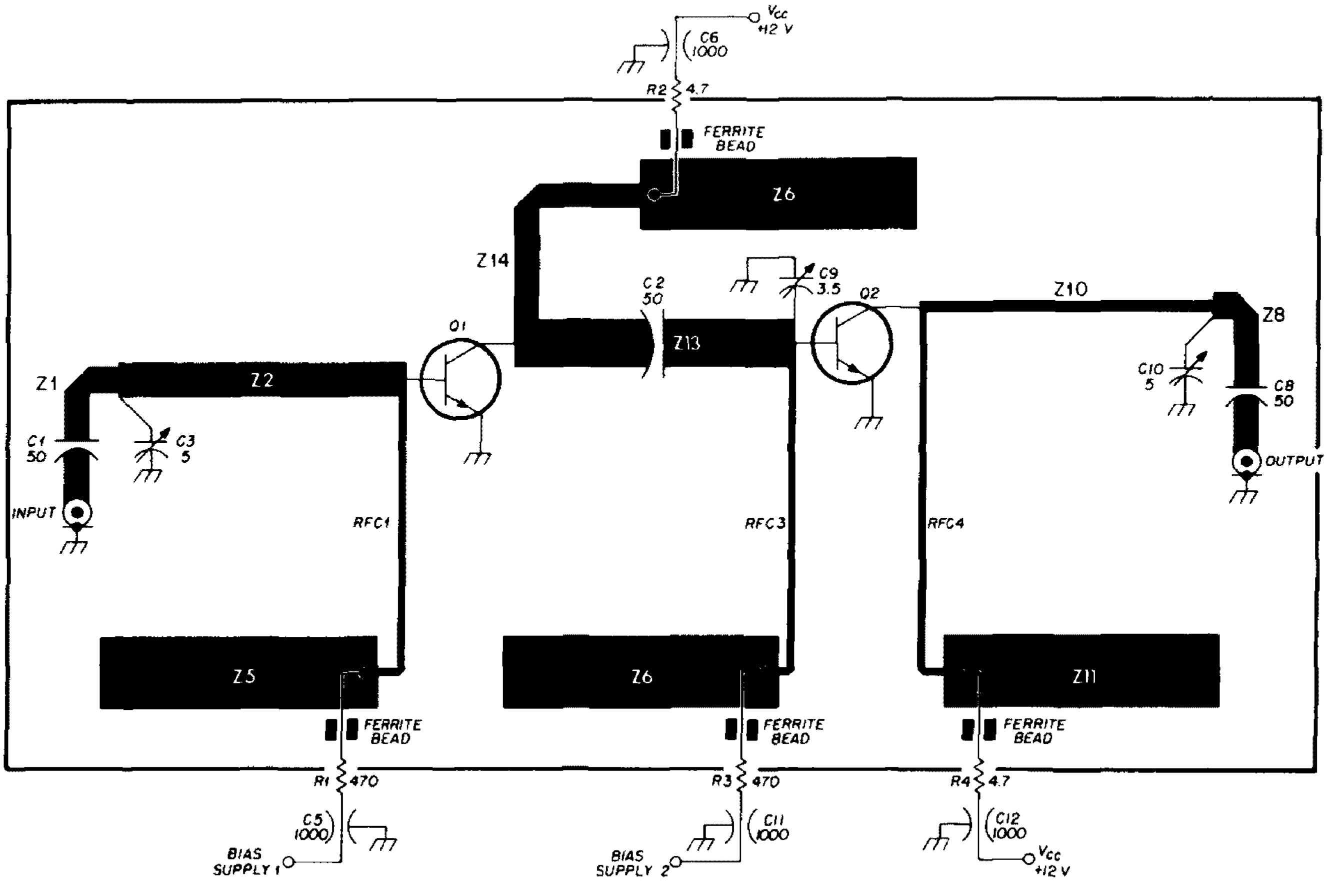


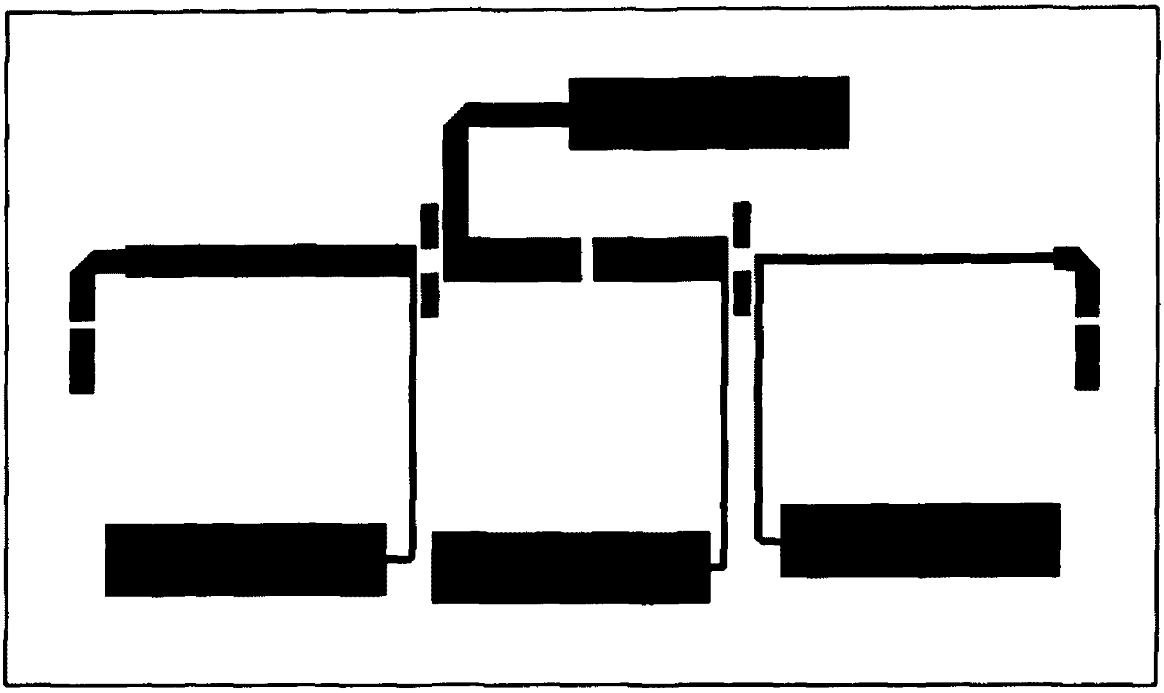
fig. 14. Modified interstage circuit for building the two 1296-MHz amplifier stages in figs. 1 and 3 on one circuit board. Circuit-board layout is shown in fig. 15.

It should be pointed out that, with two stages cascaded in this manner, it is impossible to measure the output vswr of the first stage or the input vswr of the second. Therefore, proper matching can be approximated only by tuning the amplifier for maximum gain. This is complicated by the fact that interstage matching is influenced by the output reflection coefficient, S_{22}, of transistor Q1 (which varies with I_{C1}), the input reflection coefficient, S_{11}, of Q2 (which varies with I_{C2}), and the complex impedance of the interstage network (which is controlled by capacitor C9). In practice C9 should be adjusted for maximum amplifier gain with I_{C1} set at 5 mA and I_{C2} at 10 mA. Further adjustments should then be made while monitoring total system noise figure with an automatic noise meter.

A full-size printed-circuit layout for the unified two-stage amplifier board is provided in fig. 15. Preliminary tests show the total gain of this amplifier to be within 0.5 dB of the two-module arrangement. With HP-22 transistors at Q1, both amplifier configurations yielded noise figures of 2.3 dB, measured on a Hewlett-Packard 340B Automatic Noise Meter with an AIL 7010 argon-discharge noise source.

acknowledgements

I wish to express my appreciation to George Bowden and Len Lea, both of Hewlett-Packard, as well as to Jerry Arden and Jerry Swan of California Eastern Labs, for furnishing technical assistance, device scattering parameters and sample transistors for building these amplifiers.



Z13 37 ohm, quarter wavelength microstripline, 0.18" (4.5mm) wide, 1.16" (29.5mm) long, gap in center for installing C2

Z14 150 ohm shunt inductive reactance consisting of 50 ohm, 0.2-wavelength microstripline, 0.10" (2.5mm) wide, 0.96" (24.5mm) long

fig. 15. Full-size printed-circuit layout for the two-stage 1296-MHz amplifier. For identification of microstriplines Z13 and Z14, see fig. 14.

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ham radio

digital touch-tone encoder for vhf fm

Touch-tone encoder design is simplified through the use of a new cmos tone-encoder IC — the Motorola MC14410

Jon DeLaune, W7FBB*

A new cmos integrated circuit introduced by Motorola, the MC14410, allows the design of a very compact, accurate, low power Touch-Tone† encoder system. The IC provides the full 2-of-8 encoding from a basic 1.0-MHz crystal oscillator. All that *is* required in addition to the MC14410 IC is a 2-of-7 or 2-of-8 key pad switch matrix such as a Chromerics ER-21623 or a ER-21611. Some of the outstanding features provided by the MC14410 are low power (typically 7.5 milliwatts at 5 volts during standby), full 4x4 frequency matrix (16 characters), multiple key lockout, output frequency accuracy of ±0.2%, built-in internal crystal oscillator and output available for transmitter PTT.

encoder operation

Operation is most easily followed by referring to **fig. 1**. The encoder consists of a common 1-MHz crystal oscillator, two separate frequency divider chains with decoders and a digital sine-wave generator for each of the dividers. The inputs from the key pad originate from a 4x4 switch matrix which generates a four-row (R1 to R4) and four-column (C1 to C4) input signal according to **table 2**. The row and column inputs to the MC14410 are true when they are at a logic low level. For example, when a number 4 is pressed, R2 and C1 go to a logical zero level or ground. These inputs program the low group and high group counters to provide frequencies

**Manager*, Digital *Communications* Technology Development, Computer Applications Engineering, Motorola Semiconductor Products, Inc., Phoenix, Arizona.

†Touch-Tone is a trademark of American Telephone and Telegraph.

of 770 Hz and 1209 Hz, respectively. The row inputs produce the low group frequencies of 697, 770, 852 and 941 Hz, while the column inputs produce the high group frequencies of 1209, 1336, 1477 and 1633 Hz. The row and column inputs must each provide a valid

logic exclusive NOR of the low and high group frequency dividers. The logic function is of no consequence for this particular application, but the pulsed output which is produced is important for providing a transmitter push-to-talk (PTT) signal.

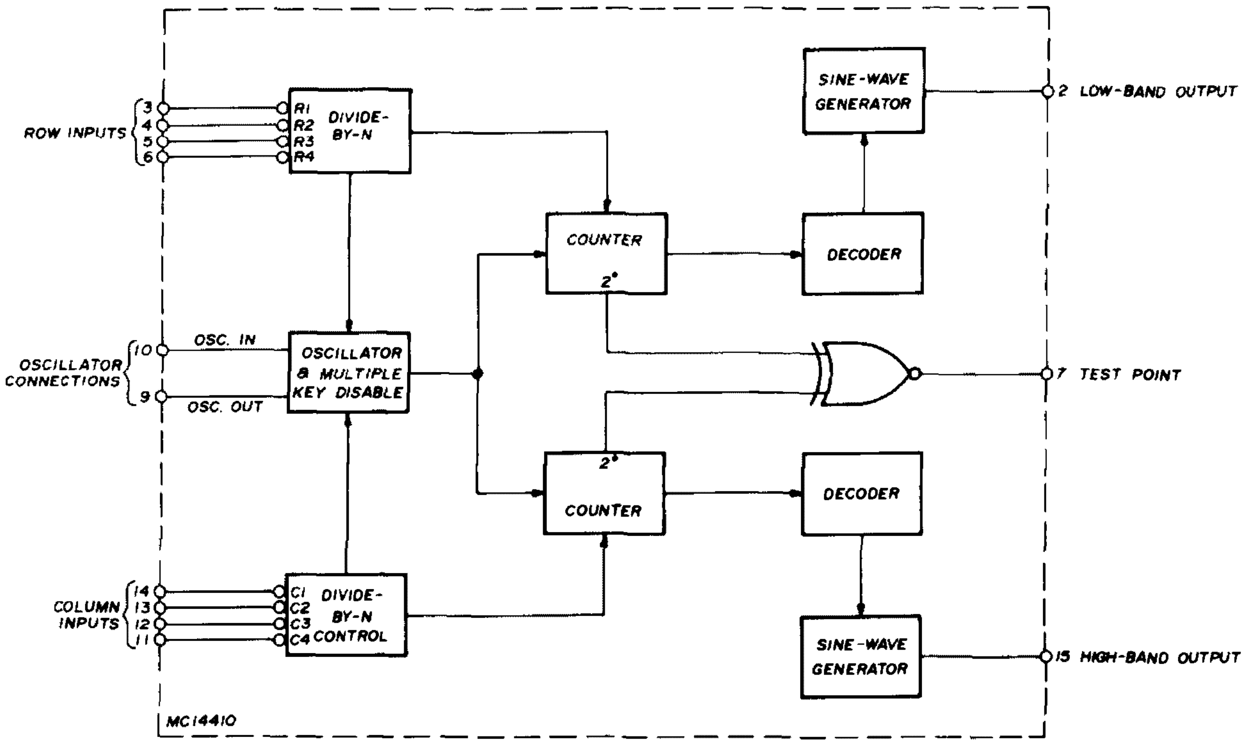

fig. 1. Block diagram of the cmos tone encoder IC, the Motorola MC14410.

1-of-4 input condition, as shown in **table 1**, before a two-tone output will occur.

The low and high group frequency dividers provide inputs to two eight-level digital-to-analog converters which synthesize the two sine-wave group tones. The outputs of these generators (low group at pin 2 and high group at pin 15), when summed together, provide the resultant two-tone output signal. Summing can be accomplished by a resistor summer with capacitive filtering required to smooth out the digital steps in the digital sine wave. This rolloff is effected by following the resistor summer with a 0.01 μF capacitor to ground.

Another output of interest is the one labeled as a test point. This output is a

The 1.0-MHz oscillator is provided internally in the MC14410; the only external parts being the crystal and a 15-megohm bias resistor. The crystal should be a parallel resonant type, maximum R_{series} = 540 ohms, C_o maximum = 7 pF, test level = 1 mW, and fre-

table 1. Input to output conditions. Row and column inputs must each provide a valid 1-of-4 input before a two-tone output will occur.

row lines depressed	column lines depressed	tone group outputs
none	none	off
none	one	off
one	none	off
one	one	both on
two or more	one	high only
one	two or more	low only
two or more	two or more	off

quency = 1.0 MHz ±0.1% with 13-pF load capacitance.

tone encoder

Now that we have a general understanding of what the MC14410 tone encoder can provide, let's look at the will drive loads down to 600 ohms with a level of 0.4 volt rms. This should be more than adequate for all 600-ohm radio systems such as produced by Motorola and others. The MPS-A17 amplifier provides a voltage gain of about five. If a level of only 0.1 volt rms

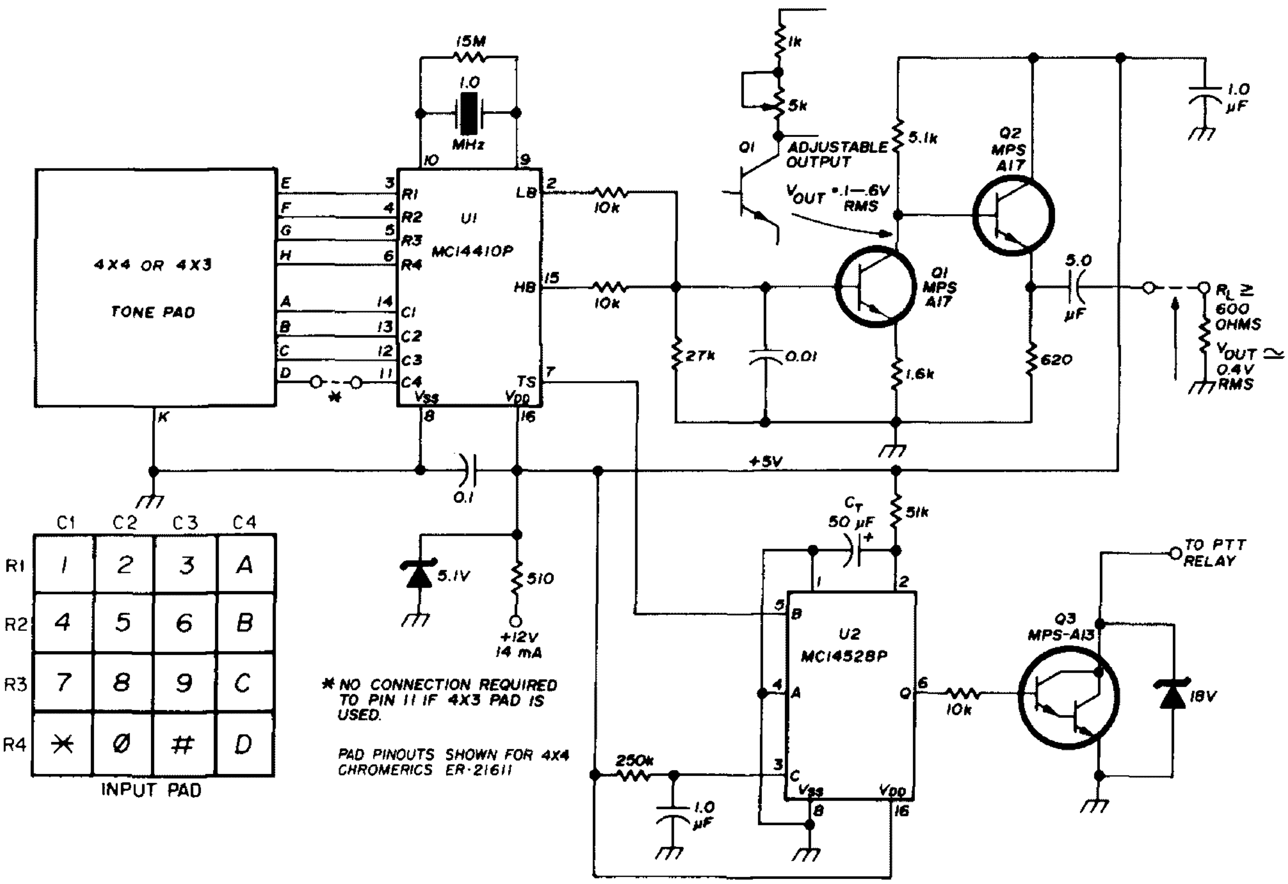

fig. 2. Digital tone encoder using the MC14410. With C_T = 50 μF, PTT hold time is 1 second (see text). Key pad matrix is Chromerics ER-21623 or ER-21611*.

amateur radio application shown in fig. 2. Included in the circuit are the Chromerics pad, the MC14410 IC encoder, U1, the resistor summer and rolloff filter capacitor, and a tone-amplifier/ emitter-follower line driver, Q1/Q2. Shown below the amplifier is the PTT one-shot timer using a MC14528, U2. Although a 12-volt supply voltage is shown, any voltage of 5 volts or more may be used as long as a zener diode is used to supply 5 volts to the ICs.

The two-tone output level from Q2 is required for proper deviation, then the amplifier may be omitted and the 10k summing resistors may be followed by the MPS-A17 emitter-follower output stage.

The tone output to the transmitter should be ac coupled through a 5-μF capacitor. The output may also be coupled through an attenuating resistor

*For address of your local Chromerics sales representative, write to Chromerics, Inc., 77 Grand Dragon Court, Woburn, Massachusetts 01801.

to provide both the required deviation level and some impedance isolation to the normal microphone input, i.e., minimum loading of normal voice audio. The 5100-ohm MPS-A17 collector resistor could be a 5000-ohm potentiometer in series with a 1000-ohm fixed resistor. This would provide an adjustable output level between 0.1 and 0.6 volt rms.

If the transmitter you plan to use with the encoder has a high input impedance audio circuit, a large isolation resistor is required between the output of the tone encoder and the audio input to the transmitter. This resistor may require values between 50k and 300k for proper transmitter deviation.

The two digital group sine waves and the final summed output are shown in **fig. 3**. The upper waveform has a frequency of 697 Hz and a peak-to-peak amplitude of 0.7 volt. The center waveform has a frequency of 1336 Hz with an amplitude of 0.7 volt peak-to-peak. The bottom waveform is the two-tone output that is applied to the audio input of the transmitter; this signal has an amplitude of about 2.2 volts peak-to-peak.

The PTT switching circuitry allows the operator to simultaneously Touch-Tone dial and key the transmitter. The MC14528 retriggerable one-shot IC has a 1.0-second hold time after tone release, i.e., the operator has at least one second of interdigit time available without transmitter dropout. This time may be adjusted to suit the operator by changing the value of the 50-μF capacitor, C_T, or the 51k resistor, R_T, or both.

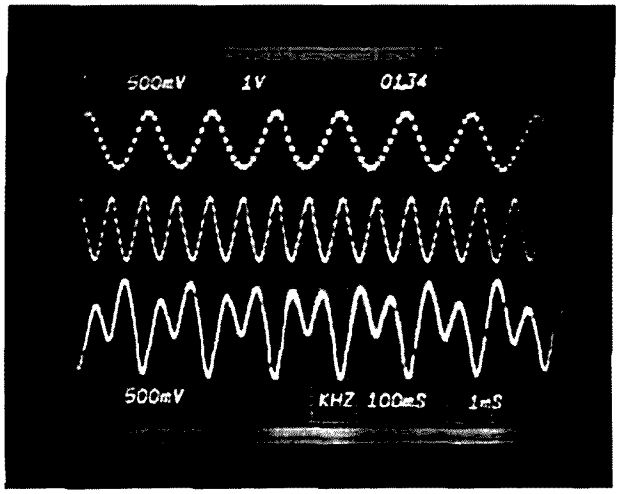

fig. 3. Output waveforms of the digital tone encoder. Upper waveform is low-group output at 697 Hz, center waveform is high-group output at 1336 Hz, and bottom waveform is the two-tone output.

table 2. Tone encoder truth table.

character input	row R1	R2	R3	R4	column C1	C2	C3	C4	high-group tone (Hz)	low-group tone (Hz)
1	0	1	1	1	0	1	1	1	1209	697
2	0	1	1	1	1	0	1	1	1336	697
3	0	1	1	1	1	1	0	1	1477	697
A	0	1	1	1	1	1	1	0	1633	697
4	1	0	1	1	0	1	1	1	1209	770
5	1	0	1	1	1	0	1	1	1336	770
6	1	0	1	1	1	1	0	1	1477	770
B	1	0	1	1	1	1	1	0	1633	770
7	1	1	0	1	0	1	1	1	1209	852
8	1	1	0	1	1	0	1	1	1336	852
9	1	1	0	1	1	1	0	1	1477	852
C	1	1	0	1	1	1	1	0	1633	852
*	1	1	1	0	0	1	1	1	1209	941
Ø	1	1	1	0	1	0	1	1	1336	941
#	1	1	1	0	1	1	0	1	1477	941
D	1	1	1	0	1	1	1	0	1633	941

ham radio

direct-reading capacitance meter

Design and construction of a simple direct-reading capacitance meter that measures 1 pF to 1 uF in five ranges

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Inflation and **scarcity** of parts have driven many hard-core homebrewers like myself to the surplus markets and swapfests where you can still find quality components at bargain prices. I have picked up a number of variable capacitors from these sources at very reasonable prices. Most of the time, however, the only way I had of determining the capacitance range was an eyeball estimate, and sometimes that proves very unreliable. I started to appreciate the idea of owning a direct-reading capacitance meter.

A brief search through back issues of the amateur magazines netted a couple of circuits, but they didn't appeal to me. It seemed there ought to be a simpler, more direct way to do it, and I began to see what I could cook up.

theory

Since the readout is a dc meter, it is desirable that there be a linear relationship between the capacitance-to-be-measured and the dc output of the measuring circuit. It so happens that the average dc value of a pulse waveform has a direct linear relationship to the duty cycle of the waveform. This is illustrated in **fig. 1**.

Duty cycle is defined here as the ratio of pulse width (in units of time) to the time between the beginning of each pulse. In **fig. 1**, T1 is the pulse width, and T2 is the period (reciprocal of frequency) of the waveform. The pulses in **fig. 1** have a peak amplitude of 10 volts; therefore, when the duty cycle is 0.2, the dc value of the waveform is 0.2 times 10 volts, or 2 volts. In like manner, when the duty cycle is 0.5, the dc value of the waveform is 5 volts, and when the duty cycle is 0.8, the dc value is 8 volts. A general expression for the

dc value of the waveform may be written as

$$\text{dc value} = \frac{T1}{T2}(E_p) \qquad (1)$$

where E_p is the peak amplitude of the pulse in volts. If T2 and E_p can be held constant, then the dc value will be directly proportional to T1, the pulse width. T2 can be held constant by making the pulse train have a constant frequency, and E_p can be made constant by regulating the power supply voltage to the pulse-forming circuit.

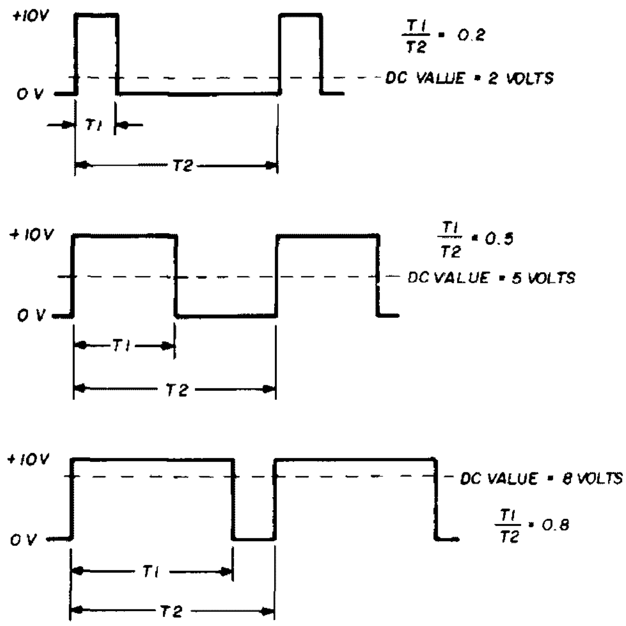


fig. 1. Average dc value of pulse waveform varies linearly with duty cycle.

This constant-voltage requirement makes an ac-operated power supply preferable to battery operation.

Now, if the width of the pulse (T1) can be made directly proportional to the value of the capacitor-under-test, the problem is solved. Fortunately, a monostable multivibrator (one-shot) can be designed so that its pulse width is directly proportional to its timing capacitor, and this completes the idea for the system.

Fig. 2 shows a block diagram of how such a system may be built. The trigger source is simply a free-running pulse generator which has a constant frequency and produces a narrow negative output pulse. Each time a trigger pulse occurs, the one-shot multivibrator initiates an output pulse whose width is determined by the capacitor-under-test. The larger the capacitor, the wider the pulse. Since a dc meter reads the average value of the pulse waveform, the meter may be calibrated directly to read capacitance. Care must be taken, however, that the pulse width does not exceed the time between trigger pulses. Also, the frequency of the trigger source must be high enough to prevent jitter of the meter needle.

circuit

A complete schematic of the capacitance meter is shown in fig. 3. The trigger source consists of a programmable unijunction transistor, Q1, an inverter-amplifier, Q2, and their associated components. There is nothing particularly critical about this part of the circuit, but the output trigger pulse at the collector of Q2 should have a frequency pretty close to 500 Hz, and the pulse amplitude should be about 12 volts (the power supply voltage). The pulse width of the trigger is about one microsecond in this circuit. Any trigger circuit which will provide this output may be used.

A Signetics NE555V timer IC is used as the one-shot multivibrator. According to the data sheet, the output pulse width of this device is given by

$$PW = 1.1RC \qquad (2)$$

where PW is pulse width (the same as T1 in eq. 1), R is the timing resistor (selected by the range switch) and C is the capacitance-under-test. Substituting into eq. 1,

$$\text{dc value} = \frac{1.1\,RCE_p}{T2} \qquad (3)$$

With the range switch in any one position, all of the terms on the right-hand

side of eq. 3 are constant except C, the unknown capacitor; therefore, the dc value of the output voltage has a direct linear relationship to the capacitance being measured.

The resistors associated with the range switch (1k through 10 meg) should have a tolerance of 5 percent or less. A 10k trimpot in series with the meter is used to make a one-time calibration; once adjusted, it needs no further attention.

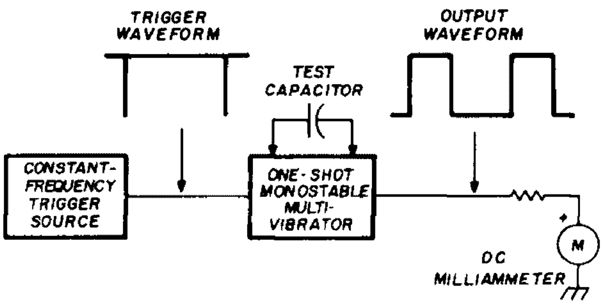


fig. 2. Block diagram of the direct-reading capacitance meter.

It was necessary to include a front panel zero-adjustment pot for the lower capacitance ranges. This is because the input capacitance of pins 6 and 7 of the NE555V and stray capacitance of circuit wiring is about 25 pF. This produces an output pulse even when there is no capacitor connected to the test terminals. Without the zero adjustment pot to buck out the voltage from this pulse, the meter will read 25 percent of full scale when the range switch is in the 100 pF position with no capacitor connected to the test terminals.

The zero adjustment circuit must have a relatively low resistance so that variations in its setting will have no appreciable effect on calibration. Since the zero setting circuit consists of a 470-ohm resistor and a 100-ohm pot in series across the 12-volt supply, it draws about 20 mA. This is another reason battery operation was ruled out.

The simple zener-regulated power supply provides 12 volts at up to 50 mA when plugged into a 117-Vac line (see fig. 4). No power switch or pilot light was included because the instrument is plugged into the ac line only when it is in use.

construction

A 5x7x3-inch (12.7x17.8x7.6cm) aluminum box houses the capacitance meter very nicely. Most of the circuitry is mounted on a 3½x3-inch (8.9x7.6cm) piece of perf board which is mounted directly to the meter terminals. The only important point in layout is to try to minimize the stray capacitance asso-

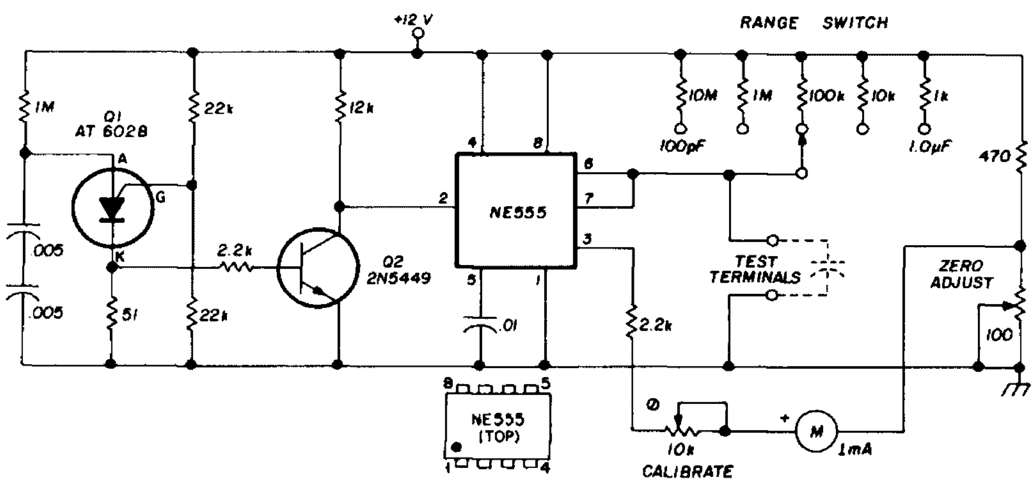


fig. 3. Simple direct-reading capacitance meter covers zero to 1 μF in five ranges, and uses easily available components.

ciated with the wiring to the range switch and test terminals. I used a Radio Shack two-terminal pushbutton strip (274-315) for the test terminals. The circuit ground should be connected to the aluminum box.

The meter is a Radio Shack 0-1 milliammeter (22-052) which has a scale

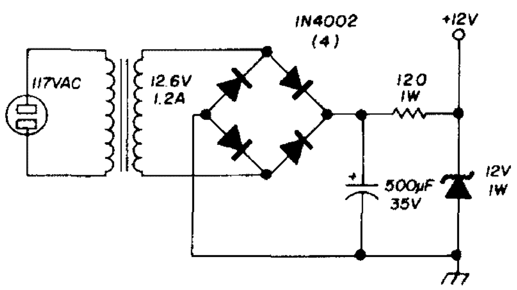


fig. 4. Simple power supply for the direct-reading capacitance meter.

length of about two inches. When the range switch is in the 100 pF position, the meter reads zero to 100 pF with minor scale divisions corresponding to 2 pF. It is not difficult to resolve a change in capacitance of 1 pF on this range.

Nearly all of the parts are available from Radio Shack. A notable exception is the 100-ohm zero adjustment pot, but any 100-ohm pot should work okay, and several are listed in the Allied Electronics catalog.

calibration and operation

Plug the ac power cord into the power line and set the range switch to 100 pF. With no capacitor connected to the test terminals, adjust the zero knob for a meter reading of zero. Now connect a 100 pF, 5 percent mica capacitor to the test terminals, and adjust the 10k calibration trimpot for a full-scale meter reading of 1.0 mA. This completes the calibration.

Each time the range switch is set to a new position, it will be necessary to readjust the zero knob *before* the capacitor-under-test is connected to the test terminals. This is a bit inconvenient, but I judged it a reasonable tradeoff for the simplicity of the circuit. Most ohmmeters require this type of operation.

If desired, a two-pole range switch may be used, and the extra pole used to select one of five preset 100-ohm trimpots instead of the one 100-ohm pot I used. The five trimpots would then be adjusted for a zero meter reading on each range with no capacitor connected to the test terminals. This would simplify operation by eliminating the front panel zero control but would increase complexity and cost.

Since ac is not applied to the capacitor-under-test, polarized capacitors may be tested, but care should be taken that the negative lead of the capacitor is connected to the grounded test terminal.

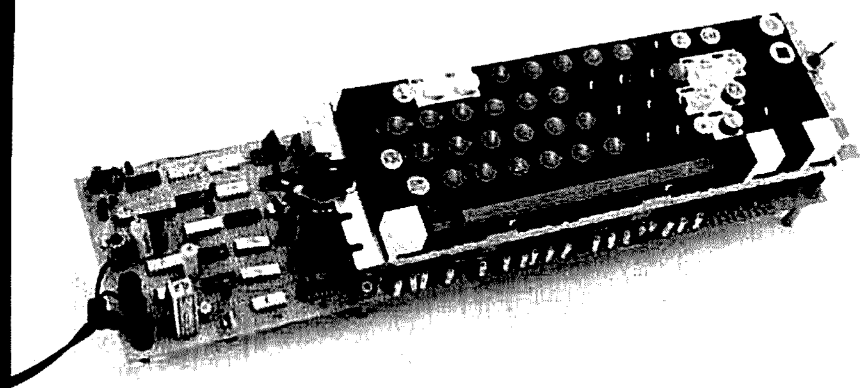
table 1. Capacitance range of the direct-reading capacitance meter.

range switch resistor	capacitance range
10 meg	0 to 100 pF
1 meg	0 to 1000 pF
100,000	0 to 0.01 µF
10,000	0 to 0.1 µF
1,000	0 to 1.0 µF

Any attempts to modify the circuit so that it will measure capacitors larger than 1 µF should be done by lowering the trigger frequency. Resistors in the range switch circuit should not have a value of less than 1000 ohms. **Table 1** shows the capacitance range that may be measured with each range switch resistor.

Since the dc voltage across the capacitor-under-test reaches a value approximately two-thirds that of the power supply voltage, any capacitor to be measured should have a voltage rating of at least 8 volts.

ham radio



keyboard morse code generator

How to build a keyboard morse generator for effortless, high-speed CW

Clarence Gonzales, W7CUU, Phoenix, Arizona 85029

Talk about one-way people! Back in 1971, Dick Vogler, K7KFA, and I developed an Automatic Fist Follower (AFF)[1] that was capable of translating incoming Morse code and recording it with a strip printer. The machine immediately increased my capability for copying Morse code by several factors, but left my fist unaffected at about 30 wpm. This incongruity proved to be embarrassing, as a fast-fisted friend in Honolulu may well recall.

I had completed development of the AFF and had tested its receipt of taped code at up to 300 wpm. All I had to do to copy code was tune my receiver to provide an 800 Hz beat note or so from a signal, lock onto the sender's fist by a quick setting of the adjustable rate control, and then set the machine to automatic copy. From then on the machine would automatically track the incoming transmission, creating a printed tape similar to that of a stock market ticker.

I was confident the machine could copy anyone on the air and must admit I entered the first live demonstration with the suppressed glee of a drag-strip driver who is carefully concealing a Corvette engine inside a stock Volkswagen chassis.

I methodically monitored the airways, awaiting suitable game. Eventually I found two operators chewing the rag in the 55 wpm range and decided to toss out the bait. The conversation I broke into was between a KH6 in Honolulu and a W6 in California.

Once I got the go-ahead I settled into an easy 20-wpm transmission, complimenting them on their fists. I then casually asked them how fast they could send.

"How fast do you want to receive?" inquired the KH6.

With appropriate confidence I again asked, "How fast can you send?"

Obviously intending a quick brush-off,

the KH6 switched to his keyboard and sent me a burst at about 75 wpm.

The AFF read him letter perfect, encouraging me to further bait him with, "Quite good, but please QRQ!"

I knew his second burst at about 85 wpm would be approaching his keyboard limit, so figured another QRQ would be the *coupe de grace*. The fellow was well prepared, however, shifting into third gear and sending me the next transmission in excess of 100 wpm. This caught me quite by surprise and though my machine was experiencing no difficulty, I began to have some doubts. Neither of us knew what was up the other's sleeve, but he must have known something was phony in Phoenix when I was receiving 100 wpm easily while struggling to transmit 30 wpm.

"Hey," I admonished, "You're not sending by keyboard or keyer, KH6."

Came the equally perplexed reply, "Well you're not receiving by ear, either, buddy!"

At this point I disclosed my 'Vette engine and he admitted to the use of an automatic McElroy keyer. We then did a little more testing during which I successfully received him at 150 wpm, his limit since he had a dirty keyer head. Later that month, after he had cleaned the head, I copied him at 200 wpm and knew I had a going machine.

Most people knew I had to be pulling a fast one, but only inbound, mind you. There were suspicions that I was taping the incoming CW, but even though my responses were relatively slow, they weren't delayed. "Very suspicious," they said.

At this point I began an avid search for a CW keyboard. I was determined to increase my sending speed, if only to keep my AFF from developing egomania.

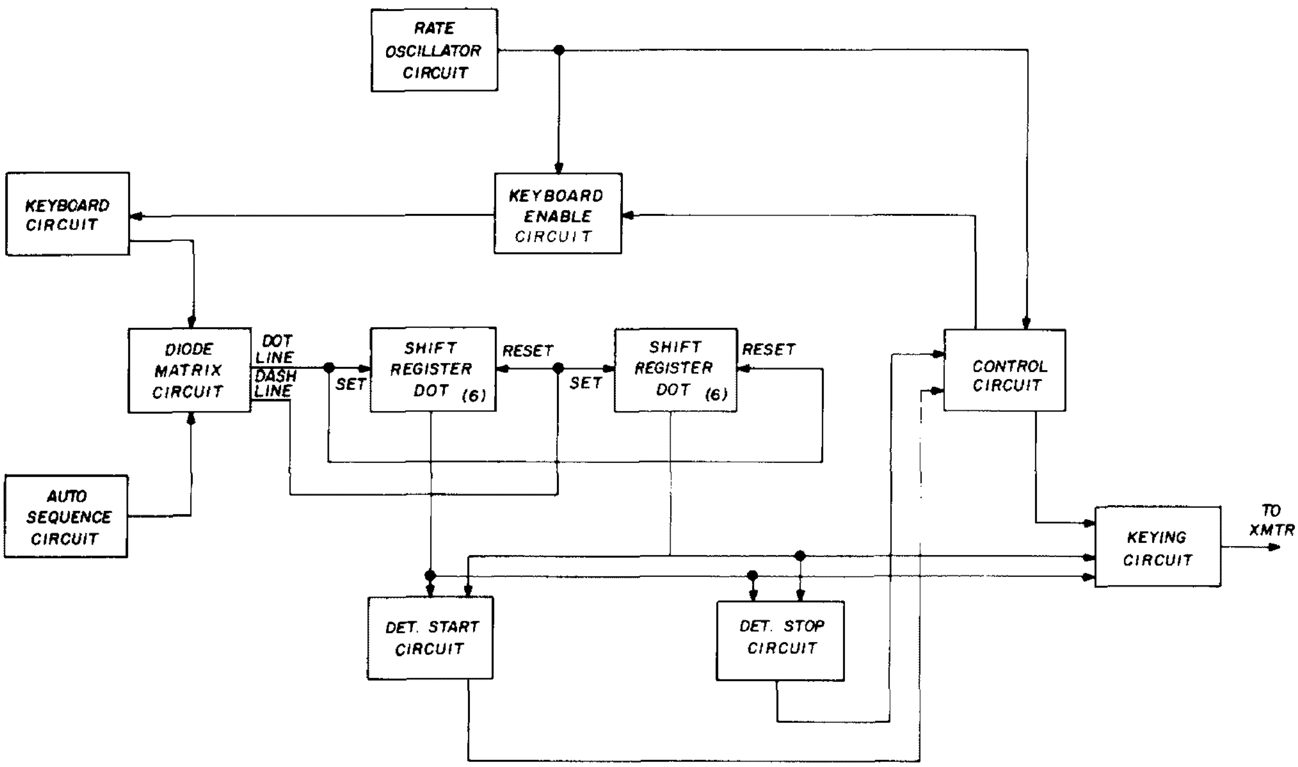


fig. 1. Block diagram of the keyboard Morse code generator. Detailed block diagram of element storage and space generation is shown in fig. 3.

CW keyboards

Let's face it though, current commercial CW keyboards are not a good answer in view of most amateurs' already over-

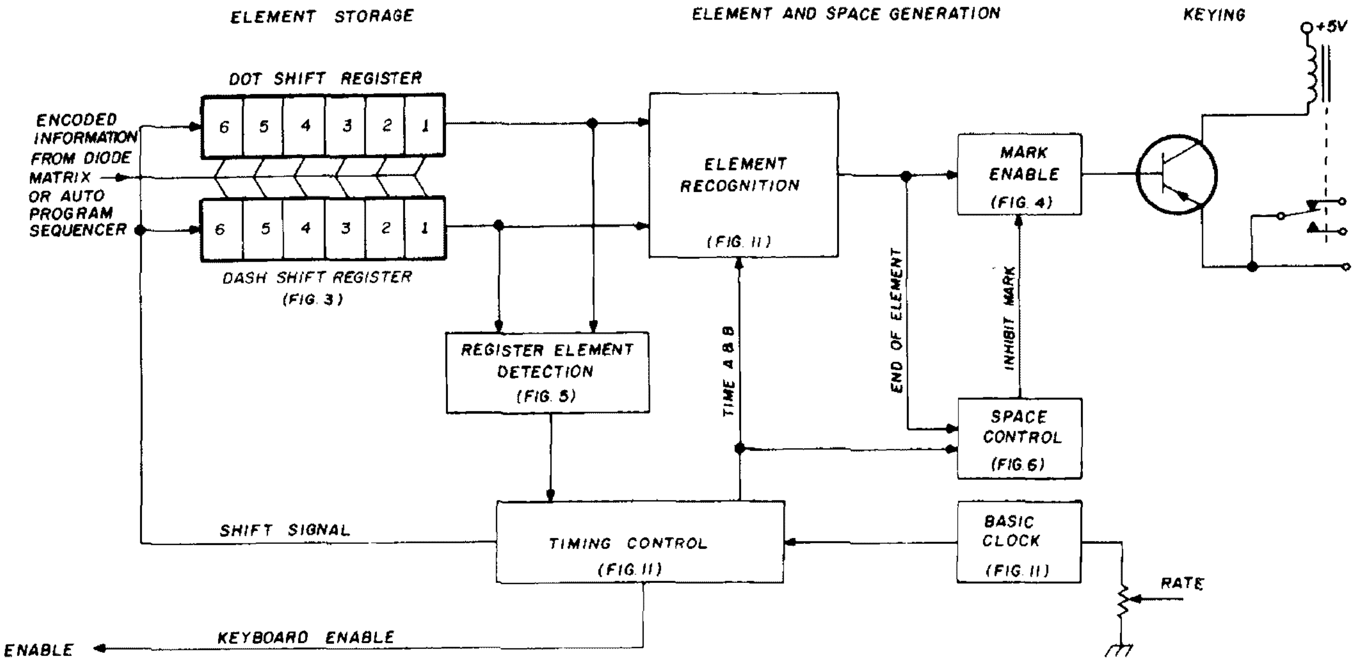


fig. 3. Block diagram of the encoding, element storage and code-generation portions of the keyboard CW generator.

extended hobby budgets. I had to find a solution which would satisfy my inherent stinginess as well as my compelling urge to "kluge," so I dug into my past issues of the various ham radio magazines and consulted my spare parts catalogs. I was convinced that I could build a better (and cheaper) CW keyboard.

I discovered during my research that code typewriters have been in existence for more than a decade.* My first reference was an article by Paul Horowitz, W2QYW describing a keyboard design,[1] in which he noted earlier attempts by others,[2,3] but he was probably the first to recognize the advantage of adapting surplus circuitry from the computer market for this purpose. This market has expanded greatly since then and today offers not only rugged, low-power integrated circuits, but also low-cost computer-related keyboards ideally suited for CW keyboard sending purposes.

*The idea has undoubtedly been around a very long time, and attempts were made to put it into practice even before integrated circuits made it practical. Such a device, designed by an engineer/ham and using only relays, was demonstrated in a relay manufacturer's booth at the National Electronics Conference in Chicago in the early 1950s. editor

Another amateur-designed automatic Morse code generator, affectionately dubbed the "Button Box" by its designer, W5VFZ, is no doubt familiar to many hams.[5] W5VFZ, while apologizing in his article for his lack of crafting expertise, nevertheless constructed an ingenious device from plywood, micro-switches and discrete circuitry. He also relied extensively upon computer surplus, adapting computer plug-in printed circuit boards containing NAND logic. His article included an "Introduction to Logic" that was quite informative for those not having the benefit of education in this area. Most of the basic concepts he covered are still applicable today, even for the newer ICs. Another logic discussion, broader in nature and more up-to-date, appeared in a recent issue of *ham radio*.[6]

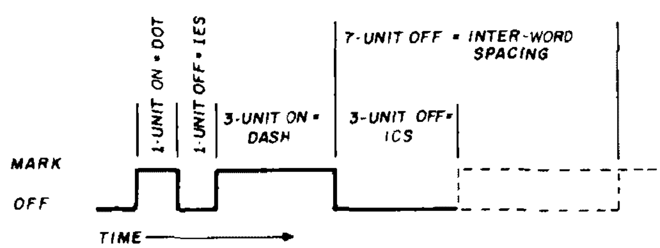


fig. 2. Morse Code element and spacing relationships. IES indicates inter-element space; ICS indicates inter-character space.

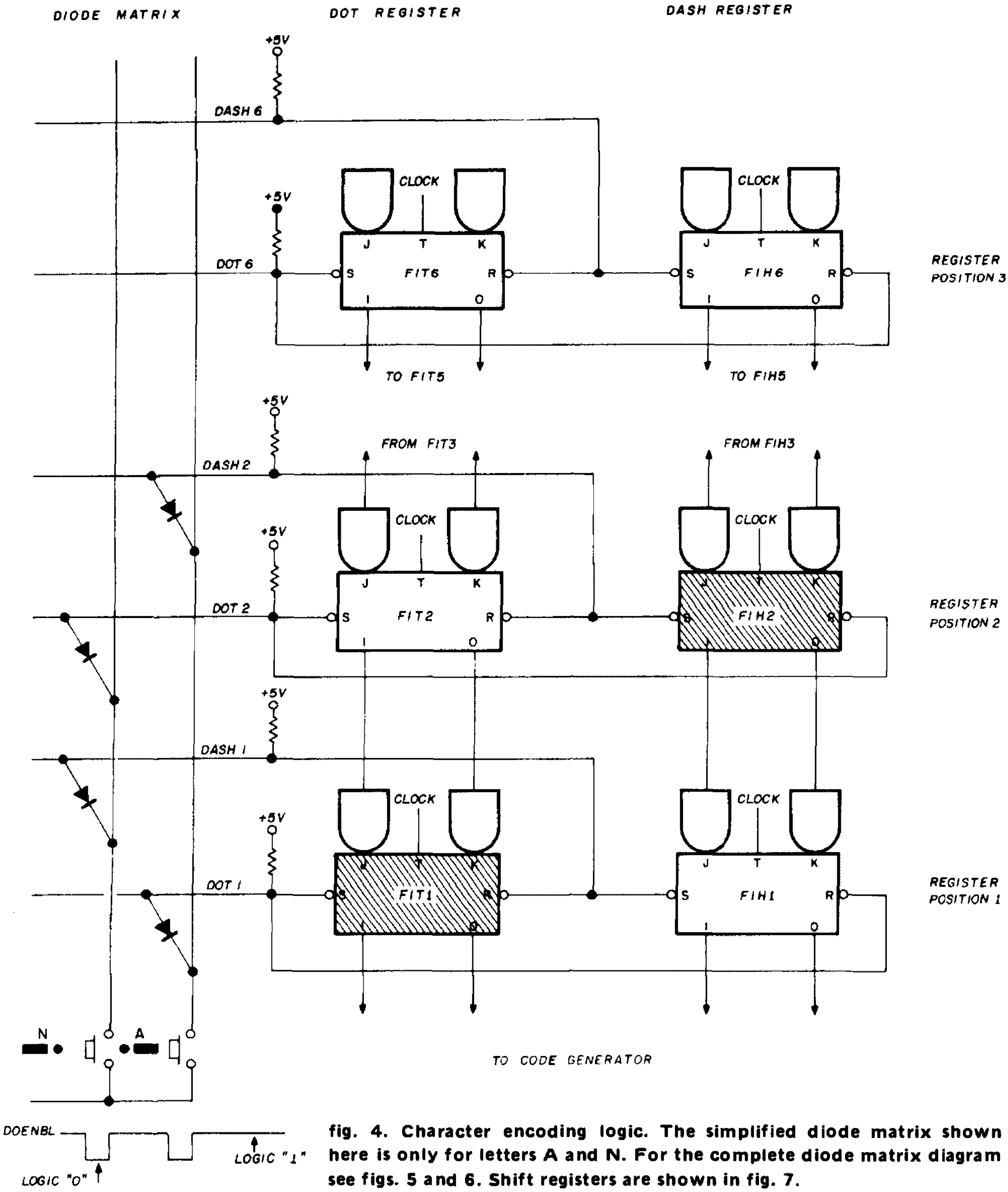


fig. 4. Character encoding logic. The simplified diode matrix shown here is only for letters A and N. For the complete diode matrix diagram see figs. 5 and 6. Shift registers are shown in fig. 7.

theory of operation

The Morse generator I developed contains a keyboard with standard-usage alphanumeric characters and *pro-signs* (pre-programmed words or messages). Upon selection of a key, the generator's circuit provides automatic encoding and transmission of that key's corresponding code elements. To do this the generator must recognize each specific key depression and translate this recognition into corresponding Morse code elements according to standardized time relationships.

Morse code characters are, in fact, distinguishable by the time relationships of the two possible states of CW transmission; *mark* or *off*. Standardized code relationships specify that a dot element is characterized by a mark-state of a single unit length and that a dash element be a

mark-state of three unit lengths. The *off* condition is used to specify separation of code elements, as well as separation of characters and words within a message. Specifically, as shown in fig. 2, a single unit space is used to separate elements within a character or *pro-sign*. Character separation is established by an off-state of three units, and a seven-unit off-state indicates a separation between words.

A single time unit does not specify any finite time period; instead it provides a time base upon which code elements and spaces can be distinguished. Single time units vary according to sending speed, but the relationships of elements and spaces to the unit length remain constant. Since I wanted to build a code generator with a flexible transmission rate, it was necessary to build in the capability of varying the basic unit timing while retaining the code and space relationships.

The following description explains the timing control necessary for varying unit timing and describes the overall encoding and generation operations involved in transforming the key selections into automatic code transmissions. This description is supported by the basic block diagram, fig. 1, and other supporting diagrams referenced through the remaining text.

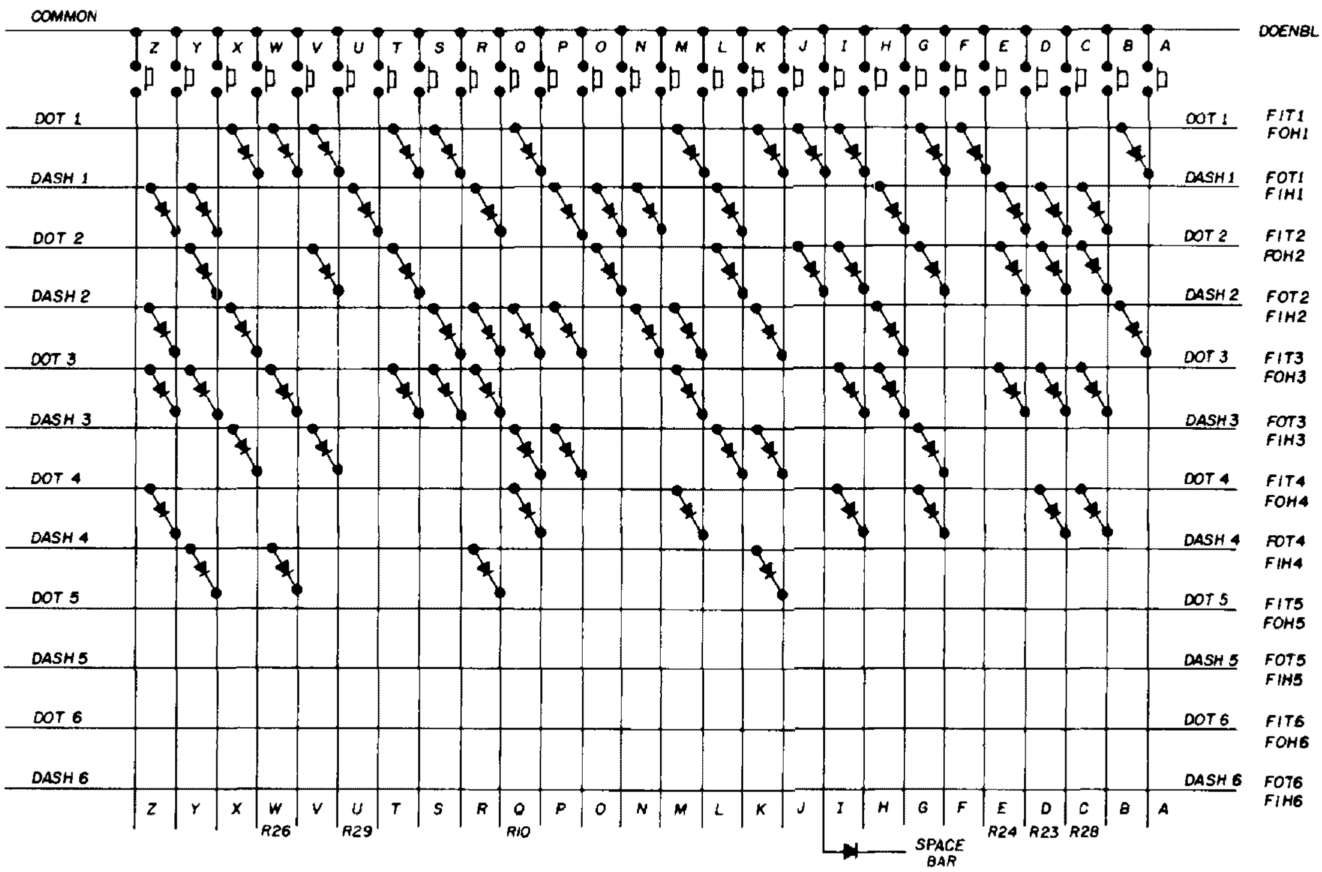


fig. 5. Diode matrix for the 26 letters of the alphabet. Diodes are 1N914 or equivalent. Connections at right side go to shift registers shown in fig. 7.

As indicated in fig. 1, the code generator's operation can be considered in two basic parts. One part is the *character encoding*, which responds to a key selection by encoding it into the selected character's dot and dash elements. This is accomplished through a diode matrix that transfers the encoded element status to a storage area. Once stored, the elements are acted upon by the second part, the *code-generation* circuitry.

The code-generation circuitry individ-

ually examines the code elements in storage and then enables the keying control circuit until the mark-state has represented the detected element. Following each element transmission the circuit automatically disables the keying control for one time period to separate the completed element from the next. After all elements for the selected key have been examined and generated, the circuit again disables the keying control for three time units before allowing the next key selection to be processed. This provides the necessary inter-character spacing.

character encoding

As shown in **fig. 1**, two shift registers perform the holding function for those elements encoded by the diode matrix when a key has been pressed. This pair of registers — one for dot elements and one for dash elements — can be considered as a functional unit representing six element positions. Six positions are necessary to fully represent the largest possible number of code element combinations. Each position has a pair of flip-flops associated with it. During encoding the lowest-order flip-flop pair is loaded with the first element to be transmitted.

If the encoded element for any specific code position is a dot, the dot flip-flop at that position of the register is set during the key depression. The dash flip-flop at the same position sets (and the dot flip-flop resets) if the element is a dash. Codes of less than six elements leave unused higher-order positions of the registers unfilled following the encoding. An unfilled position is indicated by both dot and dash flip-flops reset.

Note, for example, the shaded positions in the shift registers shown in **fig. 4**. These positions represent the flip-flops that would be set following the encoding of the letter A. Beginning at the lower-order position (one), the register could then be interpreted as containing the code for an A; a dot followed by a dash. The necessary inter-element spacing is not held

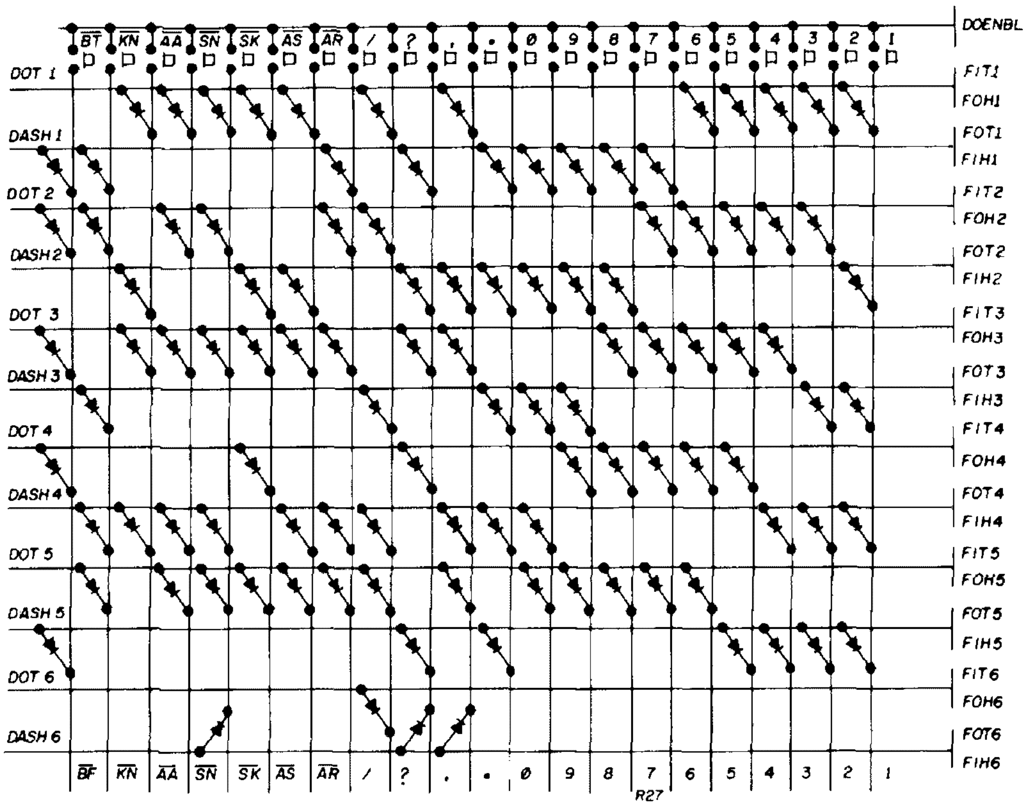


fig. 6. Diode matrix for numerals and special characters. Diodes are 1N914 or equivalent.

by the registers, but is supplied by the circuit during the actual code transmission.

Fig. 4 provides a partial illustration of the diode matrix and dot/dash shift registers used in the encoding operation.

causes the F1H1 and F1T2 flip-flops to be dc cleared. Since position one of the registers is the first examined during code transmission, this correctly represents the code for an **A** – with the exception of

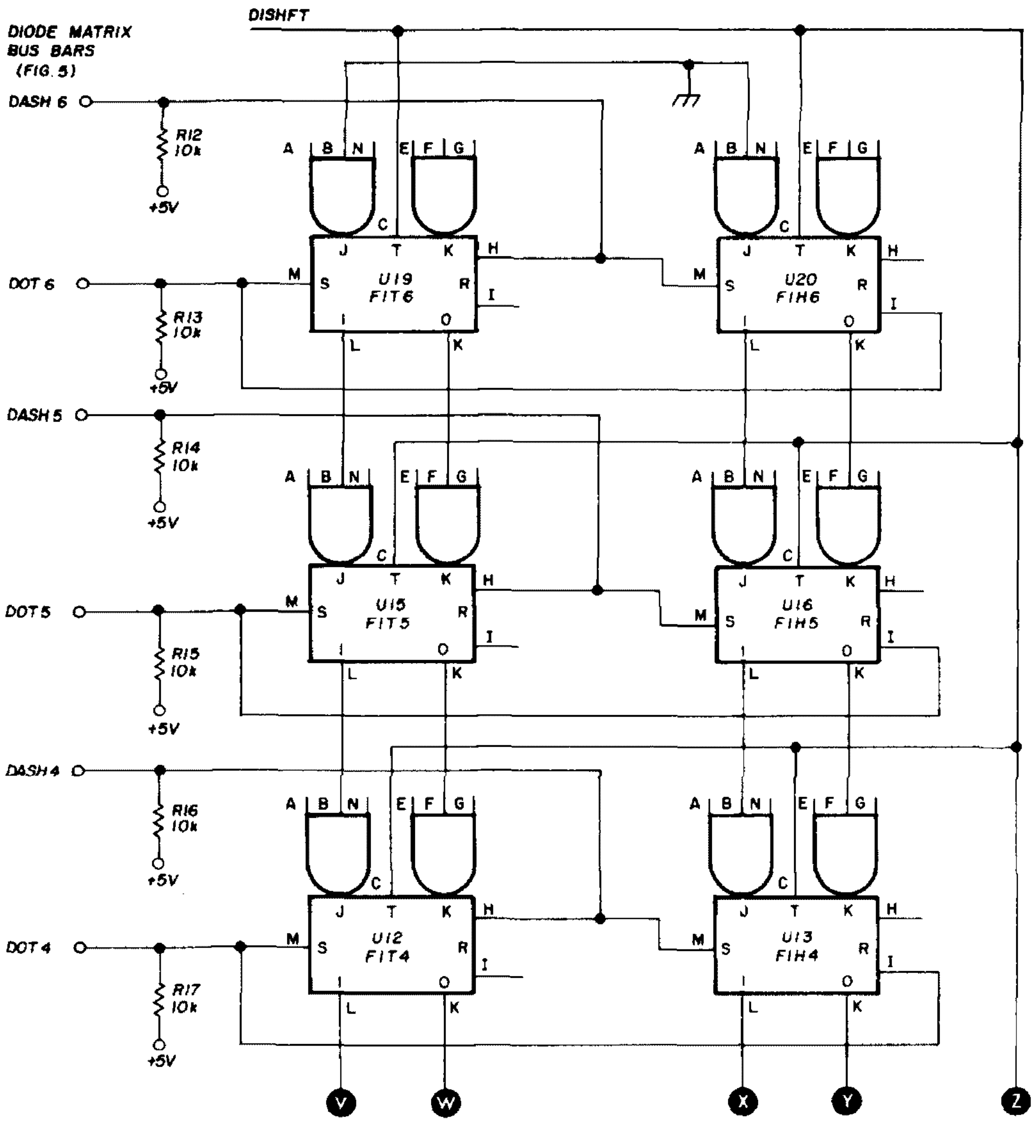


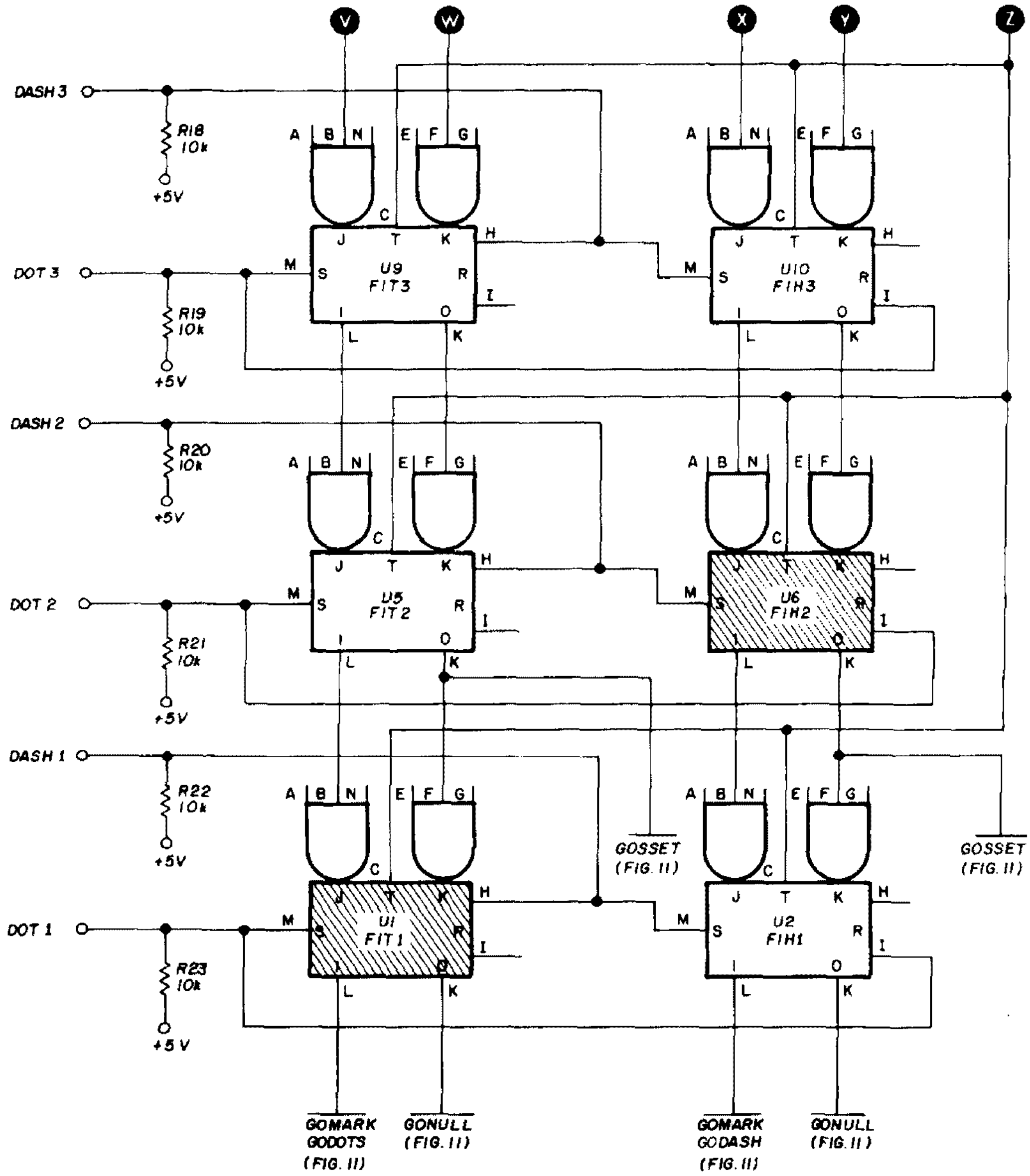
fig. 7. Dot-dash shift register receives inputs from diode matrices shown in figs. 5 and 6. All ICs are Sylvania SF200.

Only three of the six (total) flip-flop pairs and two of the possible key encodings are shown. Note that selection of an **A** permits the clock-driven enable (*doenbl*) signal to assume a low level during each of the generator's clock pulses. This action causes the dc setting of flip-flops F1T1 (*dot register, position one*) and F1H2 (*dash register, position two*); it also the inter-element spacing, which is provided by the code generation circuit.

code generation

Following code storage the elements are sequentially shifted through the registers from the higher-order toward the lower-order positions. The lowest-order position (*one*) is examined after each

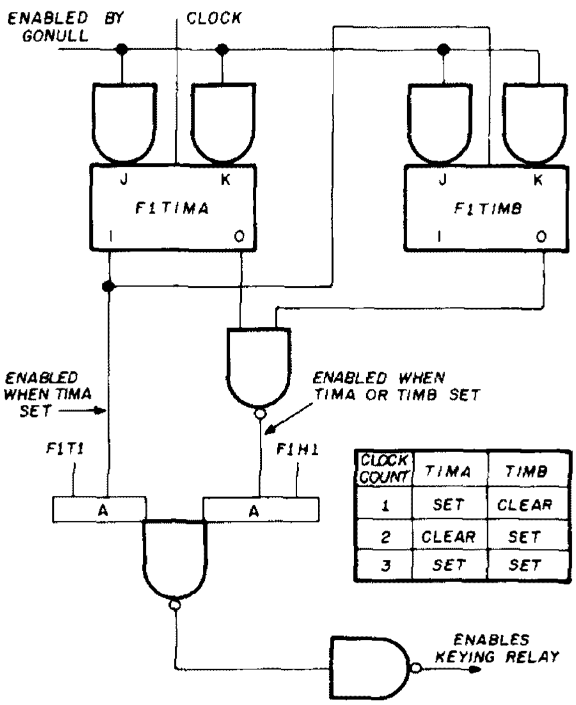
shift and the mark circuitry within the generator is enabled for the time period necessary to reflect the detected dot or dash element. **Figs. 7** and **11** depict part of the logic necessary to shift elements from F1TIMA to the *ac set* input to F1TIMB, which forms a two-stage binary counter. This permits the flip-flops to count elapsed clock periods fed into F1TIMA and thus measures time durations when



the register and to determine the correct mark period for the elements as they are examined at position one.

The timing control of the mark period is established by flip-flops F1TIMA (time A) and F1TIMB (time B), shown in **fig. 11**, which are enabled for counting during each code transmission sequence. Counting is achieved by the cross-coupled connection from the **1** output of enabling the mark circuit. The mark circuit is enabled for one clock period to represent a dot element and for three clock periods to represent a dash element. During a code character generation the mark circuit is disabled for one clock period following the mark period and the counter is readied (F1TIMA and F1TIMB cleared) in preparation for the next element.

The count versus TIMA/TIMB (time A/time B) table in **fig. 11** indicates the state of flip-flops F1TIMA and F1TIMB when permitted to count for up to three clock periods. These states are monitored by gate GOMARK (*mark*) and result in the enabling of either AND gate input when counting is initialized. If counting begins with position one of the shift



CLOCK COUNT	TIMA	TIMB
1	SET	CLEAR
2	CLEAR	SET
3	SET	SET

fig. 8. Mark enable logic. See fig. 11 for complete logic diagram.

registers containing a dot, the left AND gate is enabled only during the first count. If counting began with a dash in position one, the right gate would be enabled for all three counts. Thus GOMARK is enabled for one clock period if position one contains a dot and for three clock periods if position one contains a dash. GOMARK, in turn, enables the keying-control circuit, establishing a mark state.

Once a code element has been generated it is necessary to provide the necessary inter-element spacing and to shift the register contents so that the next element is available at position one. **Fig. 9** shows the simple three-gate arrangement necessary to determine when it is time for shifting. *If the element is a dot, the shift* takes place on the clock pulse when F1TIMA is set and F1TIMB is clear, which would be the first clock pulse. Shifting for a dash takes place on the third clock pulse when both F1TIMA and F1TIMB are set.

spacing control

Inter-element and inter-character spaces are automatically inserted in the code transmission by the disabling of the generator's mark circuitry. This disabling occurs as the circuit detects end-of-element and end-of-character conditions. An end-of-character condition occurs when the mark circuitry has been enabled for the required time to represent the detected element in position one of the registers. At this time a pulse is enabled to shift the next element to position one and to clear the time counter. One time period then elapses between the clock pulse that clears the time counter and the next one that resumes the counting, thus providing a disable for the mark circuit long enough to represent the space.

Inter-character spacing is provided by a three-unit disable of the mark circuit. One of the time units is generated in the normal manner of a space following an end of element; however, detection of the end-of-character condition sets a flip-flop that causes the time period to be extended an additional two time units. This flip-flop, shown in **fig. 10**, is set when the shift pulse occurs to transfer the last element from the register and clears when the time counter detects a count of two. As indicated by the FOT2 (dot element position 2) and FOH2 (dash element position 1) inputs to the *set* NAND gate in **fig. 10**, the last element detection actually detects when position two of the register is empty. This condition occurs when the one, which still has an element present. The flip-flop remains set, disabling the inputs to GOMARK and, consequently, the mark circuit, until the NAND gate generating the flip-flop-clear signal is enabled. It is enabled by a clock pulse during a count of two in the time counter when the registers are empty of elements.

The time counter is permitted to count for the three-unit period following a register empty period, but is then disabled. At this time the keyboard is again enabled to accept a new character entry.

automatic program sequencer

The automatic program sequencer provides the capability for generating a predetermined message in response to a key selection. The unit that I built incorporated my call sign and CQ, but any message that suits your own needs can be substituted. The message content is determined by the wiring interconnection between the sequencer's decoder output and the generator's diode matrix encoder.

The sequencer logic is constructed in 10-character groups, with each group sequenced under the control of a pair of logic functions. The logic function pair consists of a four-bit binary counter, to sequence the character transmission, and a 1-of-10 decoder to enable the predetermined character-select lines for each step in the count.

Each decoder output is connected to the appropriate character-select line in the diode matrix. As the count progresses in the binary counter the individual decoder outputs are sequentially enabled. During the time they are enabled they select a character line into the diode matrix, which encodes the character with dot/dash elements in the manner described previously for the manual operation. Upon completion of a character, the binary counter is incremented to the next count, thus enabling the next sequential output character.

The binary counter only progresses through ten counts since that is the limit of the decoder portion of the logic pair. At the end of ten counts the logic automatically steps to the next logic pair. This action continues until all logic pairs have been accessed, completing the message. The message can be prepared by depressing the sequencer key again, while holding the key down will provide continuous message transmission.

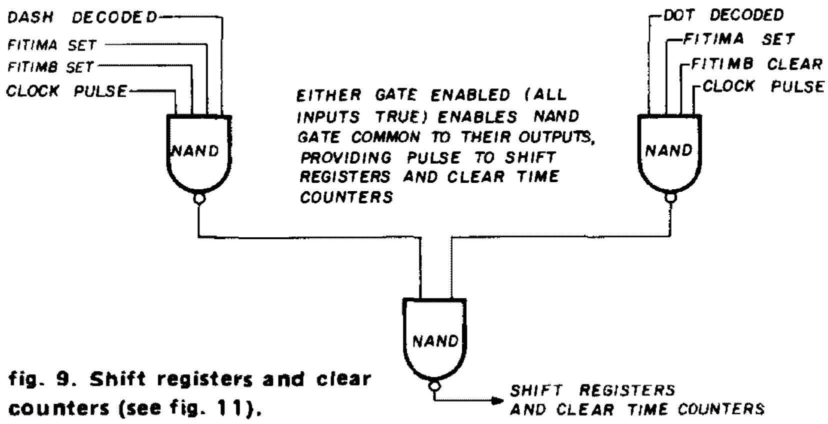


fig. 9. Shift registers and clear counters (see fig. 11).

construction notes

The entire generator was mounted on a single 4½x17-inch (11.4x43.2cm) prepunched Vector board. Approximately two-thirds of the board was used for the diode matrix layout and the few IC chips of the sequencer. The keyboard was mounted on the breadboard at four corners of its base and covers the matrix area, with the base approximately a half-inch (13-mm) above it. The other five inches (12.7 cm) of breadboard holds the remaining circuitry including the keying relay and connection jacks for the monitor and power supply.

I used a single 6-volt lantern battery to power the generator, although any supply capable of providing 800 milliamps at 6 volts would be adequate. The battery was convenient (and inexpensive), and provided me with approximately a month's service if I remembered to disconnect it when I wasn't using the equipment. There was one lapse in my thoughtfulness

when I inadvertently hooked up a 48-volt supply to the jacks. This converted my integrated circuits to disintegrated circuits and leads me to make two suggestions: First, provide a 6-volt zener at the input to serve as a shunt limiter; and second, use IC sockets if you can locate a reasonably inexpensive source.

The keyboard, which cost only $14.00,* suited my application perfectly. As you can see from the photograph, it is of modular construction, with plug-in keys and filler spaces that can be positioned along five-row segments. Using the plug-in keys and appropriate spaces you can arrange virtually any keyboard pattern you desire. I opted for the standard typewriter keyboard layout, complete with space bar, but with my special-purpose pro-sign keys added. I had to place my own labels over the keys assigned the pro-sign functions, since they are not usually-found options.

The ICs plugged nicely into the ten-hole-per-inch breadboard since their pins are spaced at tenth-inch (2.5-mm) intervals. I interconnected the ICs and discrete circuitry with 28-gauge wire, except for the power connections to the ICs which were made with 24-gauge wire. As you will note from the photograph, I stood the resistors, capacitors and diodes on end. This conserved some space and simplified their connections.

conclusion

In my previous article on the automatic fist follower, I mentioned a note that I had placed over the machine and which I later used on a post card depicting it. The note rather facetiously questioned; "Who needs to know CW?" I now have an answer – my code generator! For all practical purposes (except license exams), I could forget Morse code and not suffer.

After several months of development and testing of the generator I discovered an interesting sidelight to the machine – I was, as expected, sending faster with the generator, but I was also receiving faster by ear. This may have been due to my intense involvement with Morse code during this period, but it led me to consider the potential use of a keyboard with a monitor for Morse code instruction.

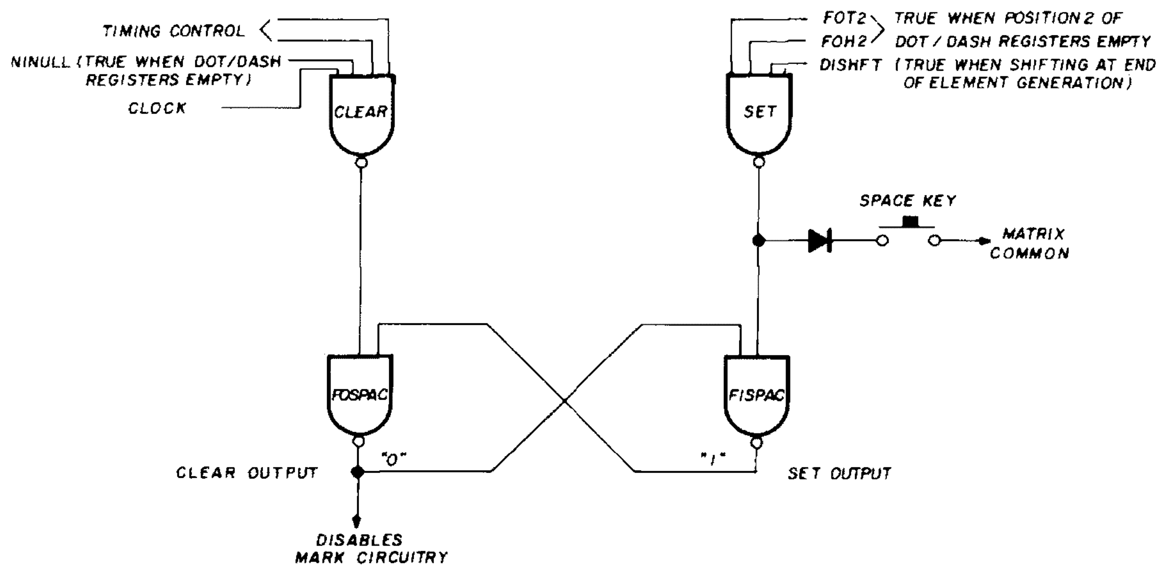
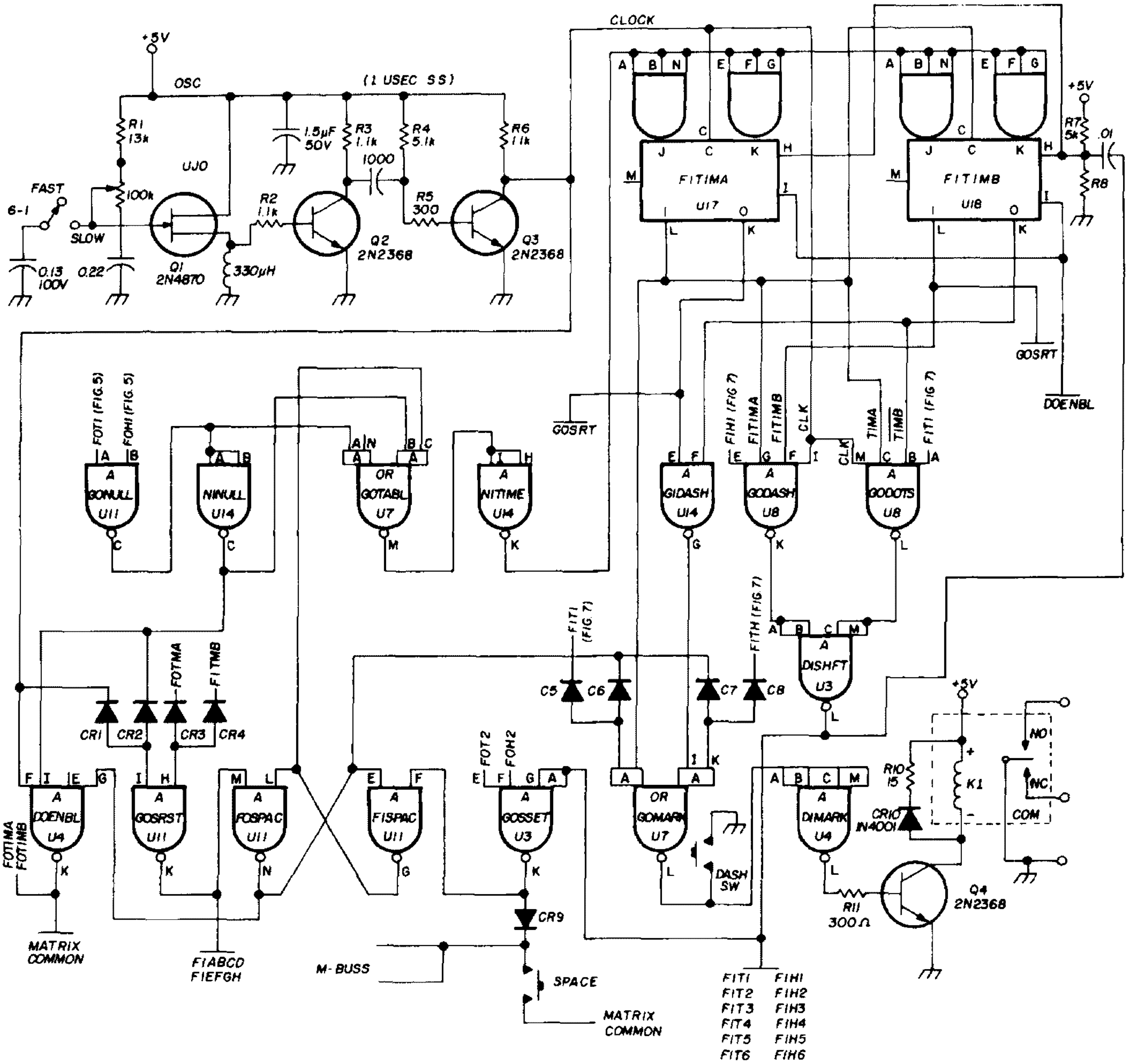


fig. 10. Space control logic (see fig. 11).

*Tri-Tek, Inc., Post Office Box 14206, Dept. H. Phoenix, Arizona 85063. Packets containing schematics, logic drawings, parts lists, theory of operation are available from VMG Electronics, 2138 W. Sunnyside Avenue, Phoenix, Arizona 85029. A similar packet is also available for the Automatic Fist Follower.



U3,U4	Sylvania SG130	U8	Sylvania SG240	U11,U14	Sylvania SG220
U7	Sylvania SG70			U17,U18	Sylvania SF200

fig. 11. Logic diagram for the timing and keying control circuitry for the keyboard CW generator. Diodes are 1N914 unless marked otherwise.

I had included a monitor with the generator since I did not intend to include a buffer to hold code while I was burst typing. This meant that I had to listen and await the completion of a character before enabling the next character. The machine locked out any characters enabled while one was being transmitted, so the potential existed of losing a character if I released a key prior to its acceptance by the logic. This characteristic required only a small amount of practice for me to accommodate, and probably trained me to be a better typist since it required a rather even typing rate. I could adjust the generator to a rate near my typing limit and operate with little or no delay.

During this period of acclimation I was forced to listen to near-perfect CW. Prior to development of the generator my comfortable CW transmission rate was a little over 15 wpm (with a somewhat "swinging" fist) and I could receive at

approximately 20 wpm. After the design and test period I found my speed had increased to approximately 30 wpm for both send and receive, with a capability of about 50 wpm in the burst mode. I also found when returning to my bug that

The question is intriguing, but probably somewhat academic in view of my daughter's reaction to the entire affair. She observed my painstakingly contrived equipment that converted a touch of the finger to a coded message received by an

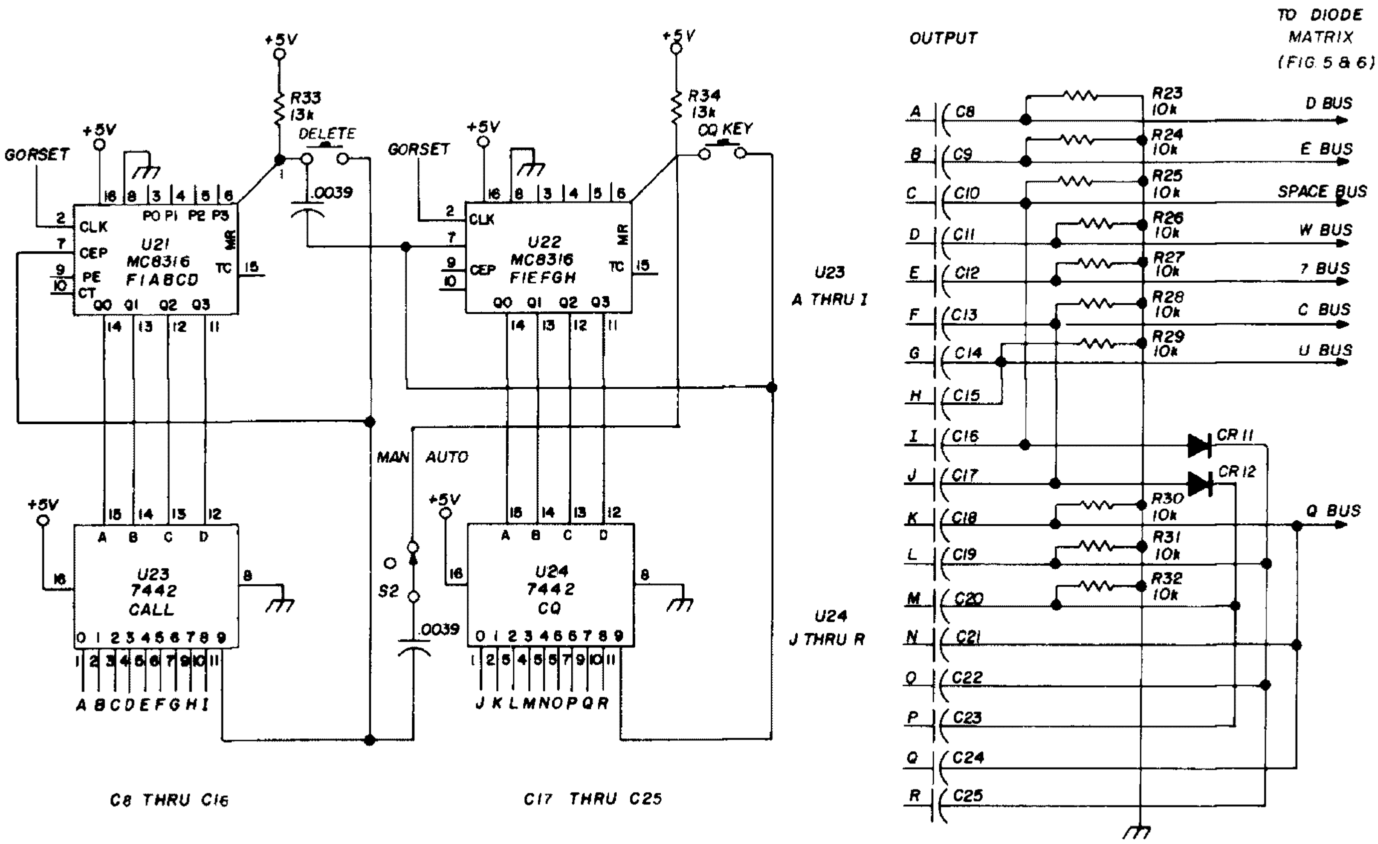


fig. 12. Automatic program sequencer for the keyboard CW generator. All capacitors are 0.0039 μF, 100-volts. Diodes are 1N914 or equivalent.

I had straightened out some of the undesirable "swing" from my fist.

I then decided to use the generator to teach Morse code to my daughter, who was completely inexperienced in both code and typing. In a short time she was capable of sending and receiving all letters of the alphabet at a rate of 5 wpm, a rate required to pass the Novice license exam. This leads to an interesting question: Would a person knowledgeable of the rules for a Novice license qualify if he appeared with and operated this equipment at the required rate? The rule specifically states that the person shall receive by ear, but an ear can be defined as a sense of hearing. Is the AFF not hearing the code when it translates it?

electronic marvel translating it into printed words. Her response? "Daddy, if you just want to see what you're typing, why didn't you just buy a typewriter?"

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ham radio

hex inverter vxo circuit

Basic data for a variable-frequency crystal oscillator plus buffer covering 2-20 MHz — all on one IC chip

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Recent experiments with TTL hex inverter ICs in variable-frequency crystal oscillator circuits (VXOs) have been quite favorable, even with simple circuits. The following text presents crystal data, circuit design using the Signetics N7404A (although its equivalent may be substituted), and performance data on VXOs covering 2-20 MHz. Simplicity, TTL compatibility, and modern design are emphasized.

device data

The following information is presented courtesy of Signetics, Inc., for the benefit of readers who may not have access to data sheets for these devices. Each inverter in the 54/74 TTL series (plastic dual-inline package) has the absolute maximum ratings (in free air temperature) shown in **table 1**. Some of the features of this logic family are presented below.

Logic definition. Logic is defined in terms of standard positive logic, in which low voltage is logical zero and high voltage is logical 1. Each input requires current into the input at a logical 1 voltage level. This current is 40 μA maximum for each emitter input.

table 1. Device absolute maximum ratings.

Supply voltage, V_{CC}*	7V
Input voltage, V_{in}*	5.5V
Operating free-air temperature range:	
Series 74 circuits	0°C to 70°C
Storage temperature	−65°C to 150°C

*Voltage values are with respect to network ground terminal.

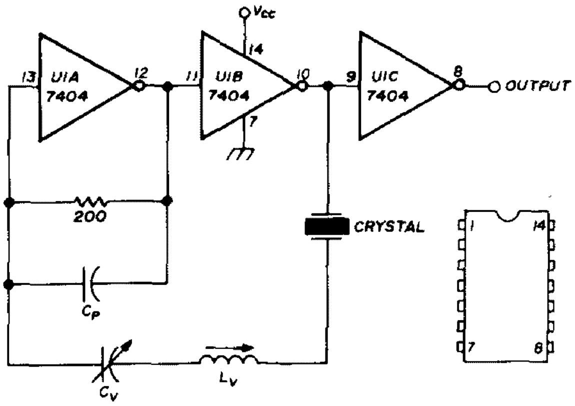


fig. 1. Hex inverter VXO using the Signetics N7404A hex inverter IC.

The switching characteristics of this IC family are typically 8 and 12 ns to logical zero and logical 1 voltage level, respectively.

Clamping diodes. All devices in the Signetics 54/74 logic family incorporate input diodes. Each clamping diode will limit negative excursions at the input to 1.5 V maximum below ground, even if −12 mA of current is drawn.

vxo circuit description

Three inverters are used per oscillator circuit (**fig. 1**). Two form the oscillator and the third is the output buffer. Positive 5 Vdc at 50 mA is connected to pin 14; negative is ground (pin 7).

Three inverters are connected in series: pin 12 joins pin 11, pin 10 connects to pin 9, and pin 8 is the output connection. A 200-ohm (linearizing) resistor is connected between pin 13 and the junction of pins 11 and 12. A frequency-limiting capacitor, C_p, shunts this resistor. The crystal is connected in series with an inductor and a variable capacitor between pin 13 and the junction of pins 9 and 10. Starting with pin 13, you should have first a variable capacitor, C_v, an inductor, L_v, the crystal, and finally a connection to pin 9.

The N7404A circuit performs in somewhat the same way as the crystal impedance meter (CIM) described below. The crystal is in series with two inverters, thus the circuit oscillates best when the crystal appears as a pure low resistance. The situation is similar to a dog biting its own tail: the two inverters are the dog (who bites harder the sharper his teeth and the stronger his bite). You guessed it. The crystal is his teeth.

test crystals

Table 2 lists the crystals used in the investigation. The parameters shown were measured on a crystal impedance meter, which is a special circuit that can tune the crystal to its series-resonant

table 2. Crystal impedance meter data on selected test crystals. Crystals E and H are designed for third-overtone operation at 19.825 and 27.255 MHz. Crystal F is for fifth overtone at 33.825 MHz. All remaining crystals are fundamental mode.

	source, type	cut	resonances (kHz) series	resonances (kHz) 32 pF anti	C_0 (pF)	R (ohms)	Q ($\times 10^3$)
A	Military FT-171-B	BT	2219.179	2219.998	19.0	14	136.0
B	International Crystal HC6/U	AT	2998.055	2999.159	3.6	22	92.0
C	International Crystal HC6/U	AT	4998.433	5000.028	5.3	4	334.0
D	Keystone Electronics FT 243	BT	5997.536	5999.495	17.2	6	138.0
E	Military CR-9/U	AT	6610.817	6612.788	5.5	40	27.0
F	Military CR24/U	AT	6763.373	6764.408	5.5	30	68.0
G	International Crystal HC6/U	AT	8996.756	8999.661	5.2	4	184.0
H	James Knight Co., channel 23 HC6/U	AT	9088.159	9089.804	3.8	40	33.0
I	RCA HC6/U from fm receiver	AT	11647.911	11654.463	10.2	3.5	82.0
J	International Crystal HC6/U	AT	14986.986	14991.926	4.8	2	219.0
K	International Crystal HC6/U	AT	19982.309	20000.020	6.2	2	59.0

frequency. This is the frequency where the crystal appears to the circuit as a pure resistor of low value. When so adjusted, the crystal is switched out of the circuit and a resistor substituted for it. When the circuit amplitude equals that of the crystal, the resistor is said to be equal to the series resistance of the crystal, since the circuit can't tell if the resistor or the crystal is in its feedback loop.

The antiresonant frequency is determined in the same way, except that a 32-pF capacitor is now in series with the crystal. The value of C_o in the table is the capacitance between the two pins that includes the holder capacitance as well as the interelectrode capacitance of the crystal. The value of Q was calculated from the above data by:

$$C1 = \frac{2(C_o + 32)(f32 - fR)}{fR}$$

$$L1 = \frac{1}{\omega 2 C1}$$

$$Q = \omega L/R$$

$$\omega = 2\pi f$$

Where f is the frequency and the other components are as shown in **fig. 2**.

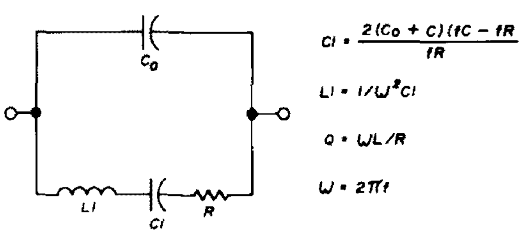


fig. 2. Equivalent circuit of a quartz crystal. The capacitance C_o is the static capacitance and is primarily a function of the electrodes sandwiching the quartz. C1, L1 and R are the equivalent elements of the motional arm.

table 3. Crystal frequency (kHz) in the circuit of fig. 1 vs shunt capacitance, C_p. L_v is zero and C_v is short-circuited.

C_p(pF)	0.0	15	100	470
A	24399.86	---bad---	2219.07	2218.49
B	8924.61	2999.37	2999.31	2999.10
C	4998.30	4998.27	4997.42	0.00
D	5997.27	5997.12	5995.36	0.00
E	19824.41	6611.33	6609.33	0.00
F	20298.39	6763.76	6761.99	0.00
G	8996.01	8995.74	0.00	0.00
H	9089.71	9089.47	0.00	0.00
I	11644.59	11642.35	0.00	0.00
J	14986.36	14984.60	0.00	0.00
K	19978.93	---bad---	0.00	0.00

Note: Bad means an erratic output; counter readings bore no relationship to the crystal under test.

table 4. VXO frequency (kHz) as a function of circuit parameters.

C_p (pF)	15	15	15	15	15	15
C_v (pF)	inf	1.0	2.0	7.8	25.5	100
L_v (μH)	0	0	0	0	0	0
G	8995.74	9011.46	9009.57	9004.17	8999.71	8997.13
H	9089.47	0.00	9099.66	9094.98	9091.70	9090.11
I	11642.35	11667.03	11665.98	11660.70	11652.80	11646.09
J	14984.60	15010.24	15006.96	14998.28	14990.11	14986.25
K	19932.75	20021.41	20012.50	20001.67		

C_p (pF)	15	15	15	15	0.0	0.0
C_v (pF)	7.8	100	7.8	100	25.5	1.0
L_v (μH)	7	7	13	13	0	0
G	9003.08	8995.33	9001.92	8992.33	8999.93	9011.16
H	9094.12	9089.33	9093.23	9087.11	9091.92	
I	11658.65	11635.35	11654.20	11592.61	11653.78	11666.94
J	14994.34	14973.37			14992.02	15009.81
K	19982.30				19992.39	20021.24

Note: A blank space means a counter output completely unrelated to the crystal — probably due to noise. A similar reaction can be obtained with a small capacitor replacing the crystal. (Counter was capable of counting to 33 MHz.)

equivalent circuit

The equivalent circuit for a crystal (fig. 2) is the parallel combination of C_o with the series string of R, C1 and L1. From this equivalent circuit it is now apparent that, at series resonance, the positive inductive reactance of L1 is

The N7404A is basically a high-frequency device. To use it for low-frequency service the high-frequency response must be inhibited by bypassing one of the inverters with a low-value capacitance. The effects of this are shown in table 3. Crystal A, you'll note,

table 5. Circuit Q as a function of VXO shifting.

	with C_v and no L_v		with C_v and L_v L_v = 13 μH		L_v = 17 μH	
C_v cap (pF)	freq (kHz)	Q ($\times 10^3$)	freq (kHz)	Q ($\times 10^3$)	freq (kHz)	Q ($\times 10^3$)
1.0	9011.22	21.7	9009.56	19.4	9007.72	16.0
2.0	9009.38	28.3	9007.17	20.4	9005.95	27.3
4.0	9007.03	49.1	9005.17	32.0	9003.33	33.5
8.0	9004.39	61.3	9001.55	49.1	8999.79	40.9
16.0	9001.14	81.8	8998.14	73.6	8994.64	54.5
32.0	8999.14	147.0	8995.32	92.0	8991.83	56.6
100.0	8997.56	147.0	8993.43	92.0	8987.69	53.0
∞	8996.76	184.0	8991.54	105.0	8983.92	64.0

Crystal tested: fR = 8996.756; f32 = 8999.661; C_o = 5.2 pF; R = 4.0 ohms

table 6. AT-cut crystals are better than BT cut for VXO applications. Here are N7404A VXO data on two-meter crystals.

series elements capacitance (pF)	inductance (μH)	frequency shift from base point AT cut (kHz)	BT cut (kHz)
∞	13	−7.271	−2.774
100.0	13	−5.715	−1.771
∞	7	−5.307	−1.324
100.0	7	−4.010	−0.835
∞	0	−3.456	−0.829
100.0	0	−2.352	−0.524
27.0	0	0.000	0.000 (base point)
7.8	0	4.327	0.575
2.0	0	8.954	0.948

Notes. (−) means below base point. No sign is above base point. At base point both are 8081.675 kHz. Other parameters are:

cut	holder	F32 (kHz)	FR (kHz)	R (ohms)	C_o (pF)
BT	FT-243	8081.649	8081.038	13	13
AT	HC6/U	8081.703	8079.103	12	4.2

exactly equal and opposite to the negative reactance of C1. The circuit is now only a resistor, R, and a capacitor, C_o in parallel, since the resistance is very low, the crystal now appears as a low-value resistor (since the few pF in C_o are swamped).

doesn't behave itself until the inverter is shunted with 100 pF. With zero pF, crystal A took off on its eleventh overtone, which would be fine if we wanted it because it was very stable. Crystals E and F are overtone crystals by design: they are tamed to their

fundamental mode with only 15 pF. The rest of the table is self-explanatory and should be studied before applying the circuit to any particular crystal situation you have in mind.

frequency shift

It makes sense to VXO only the higher-frequency crystals, since they can be moved in meaningful and useful amounts without getting into instability problems. Table 4 shows the magnitude of frequency shift you can expect from various crystals by changing the series reactances. The data for 1 pF are probably not within reach of most users, since a 100 pF variable will probably be used, and this low value is not obtainable. However, the data clearly demonstrate the value of reduced stray capacitance, since most of the easy-to-get shift comes on the low capacitance side. To shift the frequency below the fR, the series-inductive reactance must predominate.

Consider crystal J. Using an inductor of 7 μH and a 100- to 2-pF variable, the possible frequency shift is about 25 kHz above, and 13 kHz below, series resonance. With care you can VXO a 15-MHz crystal about 50 kHz. Lower-frequency crystals, of course, can be shifted less.

stability considerations

The reason we use crystals to control frequencies is simply because they are stable, high-Q devices; and when you VXO them you pay for it through the loss in effective Q in the circuit. When the VXO act is carried to extremes, you no longer have a crystal in charge of the frequency, and the circuit promptly becomes an old LC deal with the usual problems. So consider table 5 and be your own judge. A circuit Q of 20,000 is good to stay above, but it is not the end of the line. You *trade a little*, so if you are aware of what you are doing, you can get away with it under the right conditions.

table 7. Capacitor values calculated from calibration correlation for crystal impedance meter (CIM) and N7404A circuits using CIM experimental data.

experiment capacitor (pF)	predicted capacitor CIM (pF)	N7404A (pF)	correlated frequency (kHz)	experiment frequency (kHz)
1.0	1.40	1.26	9012.052	9011.123
2.0	2.11	2.09	9009.916	9009.718
4.0	3.90	3.80	9007.036	9007.145
7.8	7.43	7.40	9004.005	9004.217
32.0	32.70	33.8	8999.232	8999.181
100.0	93.20	180.20	8997.574	8997.637
∞	−705.00 (0.44 μH)	−178.00 (1.8 μH)	8996.667	8996.531

The other parameters are Co = 5.2 pF, R = 7.5 onms, and Q = 34,160. The correlation coefficient is 0.9996; the equation is:

$$\text{freq (kHz)} = 8996.667 + \frac{95.388}{C_o + C}$$

two-meter 8-MHz crystals

There are two kinds of crystals available to amateurs, the BT and AT cut. The old surplus variety *(still around) in* the FT-243 holders is most likely BT cut. The AT cut crystal is better for VXOing. However it is more expensive and is the crystal most available today from most manufacturers. In many cases crystal suppliers won't even talk to

you about the antique BT cut. Table 6 shows the amount of shift you can expect with two-meter 8-MHz crystals of both types.

calculating frequency from series capacitance

Here is a way to calculate the frequency of a VXO from the series capacitance. The N7404A IC has almost the property of a series tuned circuit without the knob; maybe you noticed this when scanning the data tables. I decided to see how close it really was and made measurements of crystal-capacitor combinations on both an honest series-resonant circuit and the N7404A circuit. The instruments used were a Radio Frequency Labs Model 459A crystal impedance meter (CIM) and General Radio Model 1612AL capacitance meter.

To get an easy-to-visualize table of comparison, I took the data obtained from both circuits and calculated a capacitor value from the correlation of reciprocal $(C_o + C)$ vs frequency, which is a straight line for all crystals. The intercept is the series-resonant frequency, and the slope is a function of the resonator cut, etc.

The line for any crystal can be determined (as it turned out) by measuring the crystal frequency in the N7404A circuit with two different capacitors in series, preferably one about 5 pF and one not larger than 32 pF. Since each crystal, even with the same cut and frequency, will be an individual with its own slope, don't expect one line to work too well for other crystals. But the same performance can be expected, so it is a useful guide. Table 7 shows that it is reasonable to predict frequency from a capacitor with the N7404A circuit. Of *interest to those of you with a counter:* you can measure low-value capacitors with a calibrated crystal (two known points are all you need if you know C_o; if not, you need three points).

ham radio

general purpose wideband rf amplifier

Design and construction of a two-stage, small-signal amplifier that may be used in the frequency range from 500 kHz to 500 MHz

Randall Rhea, WB4KSS, 1560 Jennings Way, Norcross, Georgia 30071

The basic building block of practically all electronic circuits is the amplifier. Each time a circuit is designed and breadboarded the amplifiers in that circuit must be designed, tested, troubleshot and, sometimes, redesigned. I have often thought that there had to be a better way. For many requirements the answer lies in the circuit described here. It is a standard gain block capable of operating from typical intermediate frequencies to nearly 1 GHz. It is simple, inexpensive and amplifies without oscillating!

The design goals for this gain block were high gain, wide bandwidth, unconditionally stable operation, low noise figure, simple construction and low cost. It is not possible, however, to obtain state-of-the-art performance in all these categories in one amplifier. What is achieved with the circuit in **fig. 1** is a good compromise. The amplifier will perform well as an rf amplifier, i-f amplifier or general purpose preamp.

transistor selection

The selection of transistors is very important in high performance amplifiers. The 2N5179 transistor used in this amplifier provides high gain, large bandwidth and low cost. At 150 MHz a typical device will provide nearly 14 dB gain when inserted in a 50-ohm transmission line with no tuned circuits on the input or output. Stability at this heavy loading is guaranteed. The noise figure for this device, with optimum source resistance, is about 3 dB at 150 MHz. The 2N5179 sells for $0.77 in small quantities. For further information see reference 1.

circuit

The amplifier circuit is a two-stage, capacitively-coupled, common-emitter cascade. The collector-base feedback resistors are key elements in this circuit. They provide a simple means of bias, reduce gain at low frequencies to stabilize the circuit, and lower both the input and output impedance of the amplifier. A lot of work for two resistors! Capacitor C3 decouples each stage and R5 and C5 decouple the amplifier from the supply line.

The 2N5179 transistors have a rather wide specification in dc current gain (beta). With the simple bias network used in the circuit the bias point may therefore vary from transistor to transistor, and you may find it necessary to adjust R5 so that the voltage drop

across R2 and R4 is 3 to 4 volts, each. This corresponds to about 8 mA collector current. This operating current is a compromise, with higher currents giving increased gain-bandwidth and lower currents providing improved noise figure.

The capacitors in the circuit cause the low-frequency gain to begin dropping off below about 2 MHz. Increasing these capacitors to 0.01 μF will lower the frequency response to about 200 kHz. It might be possible to use the amplifier at even lower frequencies by using larger values of capacitance, but I have not investigated the stability of the amplifier when doing this.

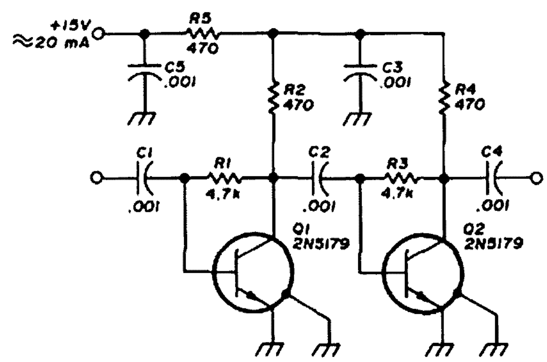


fig. 1. Schematic diagram of the wideband rf and i-f amplifier. To lower the frequency response to 200 kHz, increase the value of all capacitors to 0.01 μF. All resistors are 1/4 watt, all capacitors are ceramic disc.

construction

Mount the transistors close to the PC board, fig. 2, to keep the emitter leads very short. Be sure to ground the transistor case (fourth lead). This reduces unwanted capacitive coupling between transistor elements. If the PC board isn't used keep all leads as short as possible. In one breadboard I used no terminals, soldering each component to the next with very short leads. The result was a circuit just over one-inch (25mm) square.

A full-size printed-circuit layout is shown in fig. 3. Connectors may be

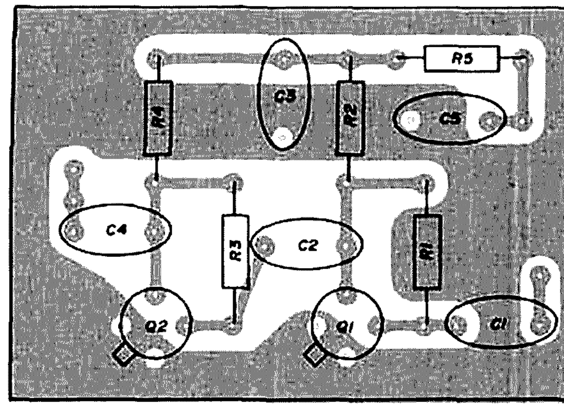


fig. 2. Printed-circuit component layout for the wideband rf amplifier. Full-size printed-circuit layout is provided in fig. 3.

soldered onto the board with short wire jumpers for the center conductor. Extra holes in the board are for these jumpers and for the +15 volt supply line. Coaxial input and output lines may be soldered directly onto the board if desired.

performance

The gain of the amplifier vs frequency is plotted in fig. 4. It is essentially the same for 50 or 75 ohms input and output loading. The measured input impedance is about 65 ohms from 500 kHz to 500 MHz with only a small series reactance. The output impedance is around 100 ohms series resistance with some reactance. The gain of fig. 4 is with 50 or 75 ohms driving and load impedance, there being little difference between the two. The maximum output level will vary some from amplifier to

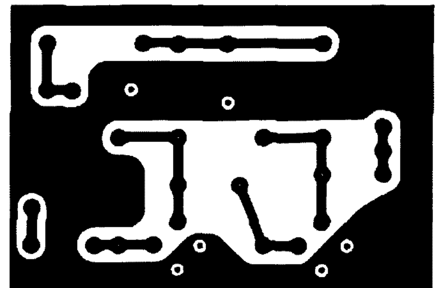


fig. 3. Full-size printed-circuit layout for the wideband amplifier.

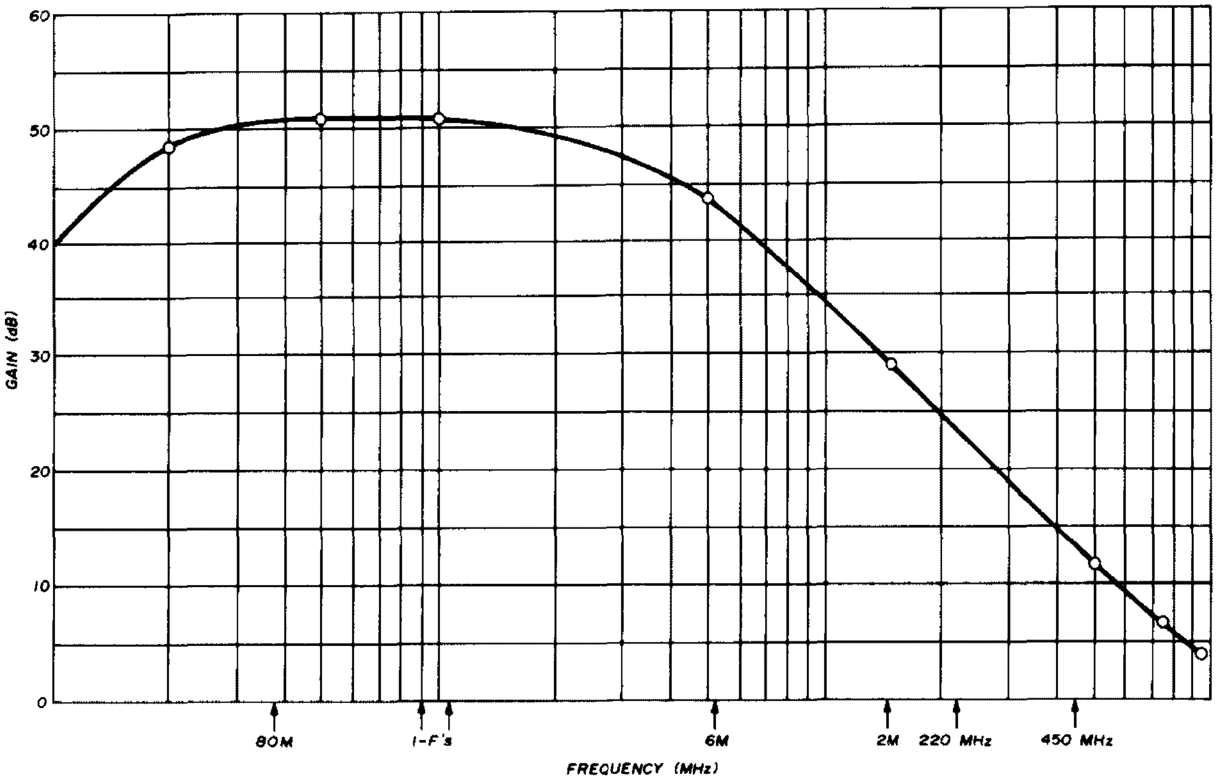


fig. 4. Frequency vs gain of the wideband rf and i-f amplifier shown in fig. 1. Circuit provides useable gain to 500 MHz and above.

amplifier because of small variations in transistor biasing. A typical value of maximum output level for 1 dB gain reduction is 300 mV.

The gain of this circuit should be treated with respect. When making connections to the board, keep the input and output leads separated. The two stages are identical so if the total gain isn't needed, connect only one stage and adjust resistor R5 for 3 to 4 volts drop across R2. I have built several of these amplifiers and none have oscillated, even with the input open, but it is a good idea to use a little caution.

tuned amplifier

Unless the amplifier will be used specifically as a broadband amplifier, tuned input and outputs should be used. This will reduce the possibility of undesired signals overloading the amplifier and will provide the desired bandpass characteristics. Since the input and output are closely matched to 50- or 75-ohm systems the tuned circuits are used only to shape the frequency response.

The Q of a tuned circuit is a measure of its 3 dB bandwidth. Consider the tuned circuit in **fig. 5**. The two 50-ohm resistors represent the generator resistance driving the amplifier and the amplifier input resistance. The total parallel loading, R, of the tuned circuit in **fig. 5** is 25 ohms. The loaded Q is given by

$$Q = \frac{R}{X}$$

where X is the reactance of the inductor or capacitor at resonance. This is ap-

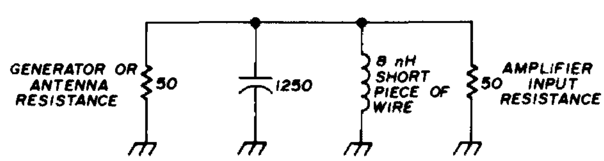
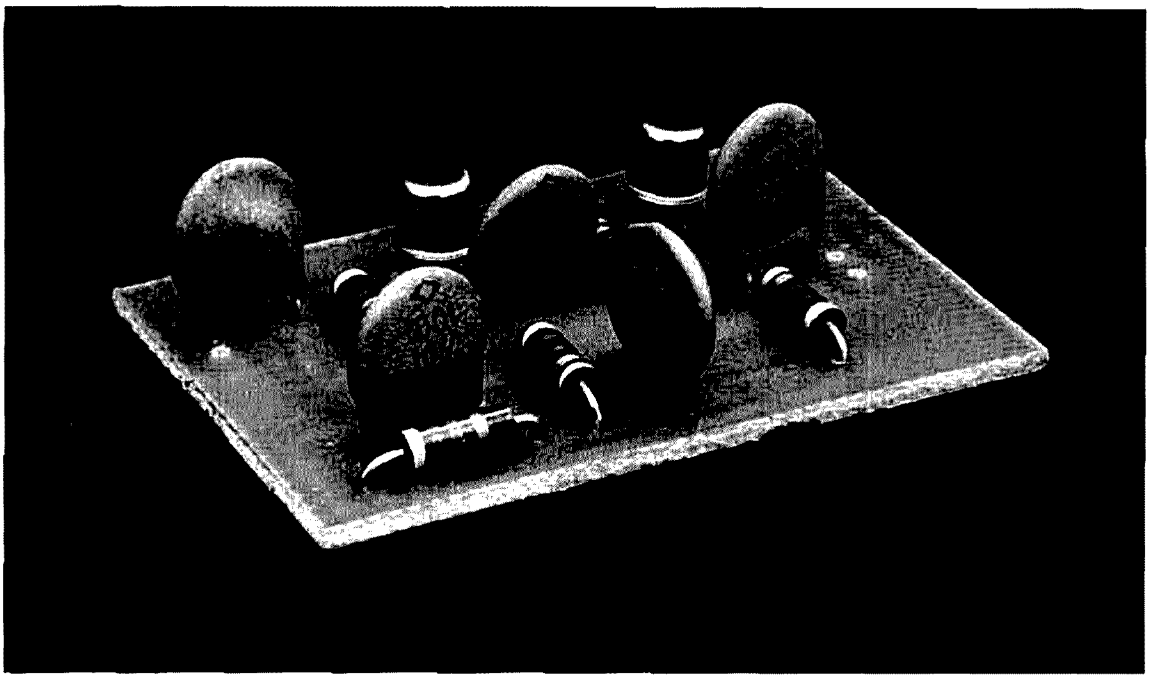


fig. 5. Model of an input tuned circuit for the wideband rf and i-f amplifier (see text).

proximately true if the loaded Q given above is lower than about one-tenth the Q of the inductor or capacitor. The Q of the circuit with no external loading is the unloaded Q. It is desirable to make the unloaded Q as much greater than the loaded Q as possible. In practice this means higher Q coils and wider loaded

To tune the circuit either the capacitor or inductor may be made variable. At high or low frequencies it might be difficult to obtain the proper reactance with reasonable values of L and C. In this case the inductor may be tapped or a capacitive divider may be used to increase the effective resistance of the



Complete wideband amplifier, shown here almost twice full size, is built on a 1.75x1.25" (45x32mm) printed-circuit board. In this view transistor Q1 is at left rear, Q2 to the right.

bandwidths. Narrowband tuned circuits with low quality coils or capacitors mean high loss in the tuned circuit. For further information on this subject see reference 2.

The resonant frequency of the tuned circuit in fig. 5 is 50 MHz. The reactance of the inductor at resonance is 250 ohms, so the loaded Q is 10. The 3 dB bandwidth is given by

$$BW = \frac{f_o}{Q_{loaded}}$$

and in this case is equal to 5 MHz. Since the inductor and capacitor should have an unloaded Q of at least ten times the loaded Q, it should be at least 100 for this case.

load. (For further reading on this subject see pages 160 and 161 of reference 2.)

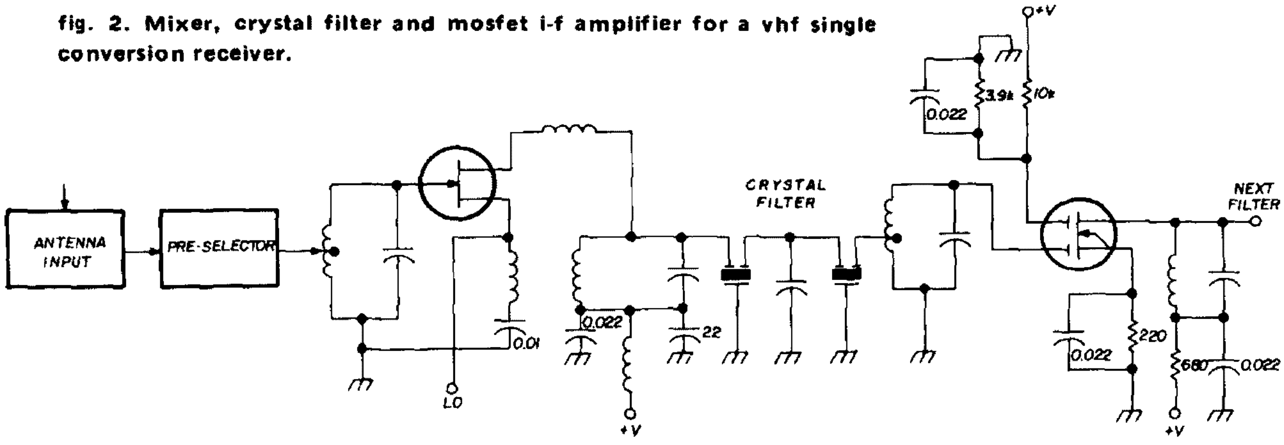
Adequate space is left at the input and output of the circuit board for compact tuned circuits. The coils at the input and output should be placed at right angles to avoid feedback caused by mutual coupling. Better still, the use of toroids will eliminate this problem and provide compact coils.

references

1. *The Semiconductor Data Library, Volume II*, Second Edition, Motorola Semiconductor Products, Inc., Phoenix, Arizona, 1973.
2. *The Radio Amateur's Handbook*, 50th Edition, American Radio Relay League, Newington, Connecticut, 1973, page 41.

ham radio

fig. 2. Mixer, crystal filter and mosfet i-f amplifier for a vhf single conversion receiver.



single i-f frequency. The structure of the monolithic crystal filter design provides high stability, reliability and excellent selectivity. Thus the single-conversion i-f can be made relatively high so that incorporated in the single-frequency i-f system shown in **fig. 1**.

The *non-monolithic* crystal filter with which most amateurs are familiar contains numerous discrete inductors,

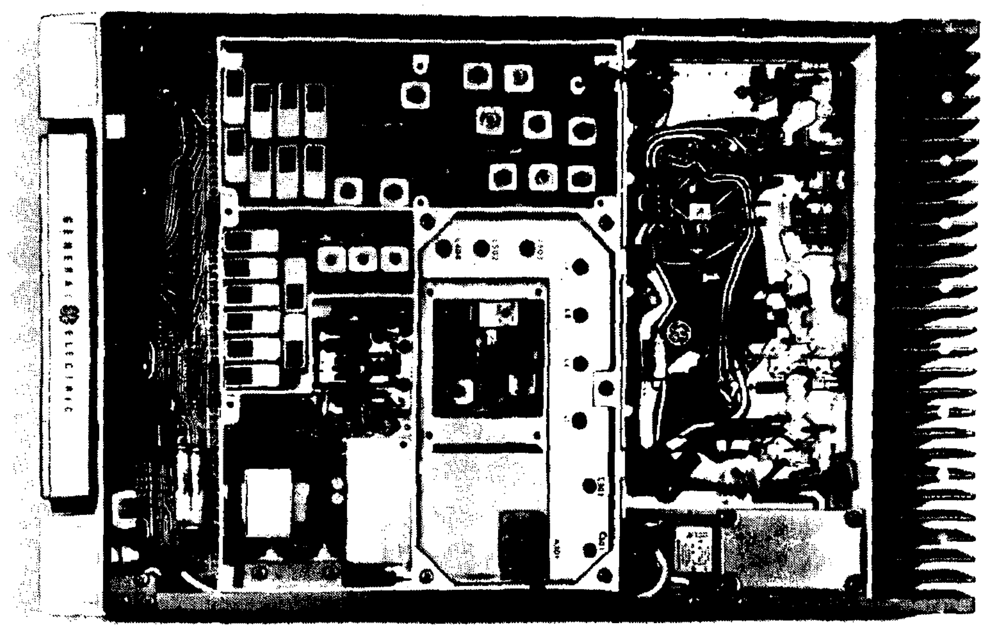


fig. 3. The GE Master II commercial vhf fm radio uses a single-conversion receiver (photo courtesy General Electric).

image response is reduced, just as it is in dual-conversion receivers with high intermediate frequencies. High-gain performance is obtained by the unique arrangement of filter and gain blocks resistors and capacitors in addition to the crystal. These components are used primarily for coupling, but they must be mounted and then carefully aligned. The *monolithic* crystal filter, however,

incorporates coupling as an inherent part of the quartz assembly (accomplished by placement of electrodes). The electrodes provide coupling and a means of converting an electrical signal frequency limit for design of such filters to accommodate a 15-kHz bandwidth is approximately 5 MHz. At the opposite end of the spectrum the practical limit at today's state of the art falls in the 25-

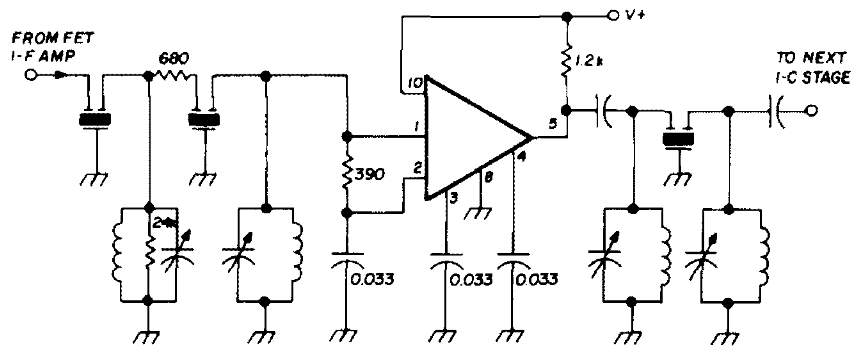


fig. 4. IC i-f amplifier with monolithic crystal filters at the input and output.

at the input to a self-resonant vibration of acoustic energy through the crystalline material. Acoustic energy is converted back to electric signal at a resonant output converter. Selectivity and coupling are controlled by the electrode geometry of the quartz blank.

Pairs and other groupings of these basic monolithic crystal filters can be assembled in modules and used to form higher-order filters. Presently the low-to 30-MHz range. (Although monolithic crystals have been made to operate to near 200 MHz, the cost is extravagant.)

Integrated circuits fit in ideally with the use of monolithic crystal filters. An integrated-circuit gain block can provide high stability with 60-dB gain and provide adequate performance up to 20 MHz or so with no instability and regeneration problems in sensible designs. Note that a single-conversion gain

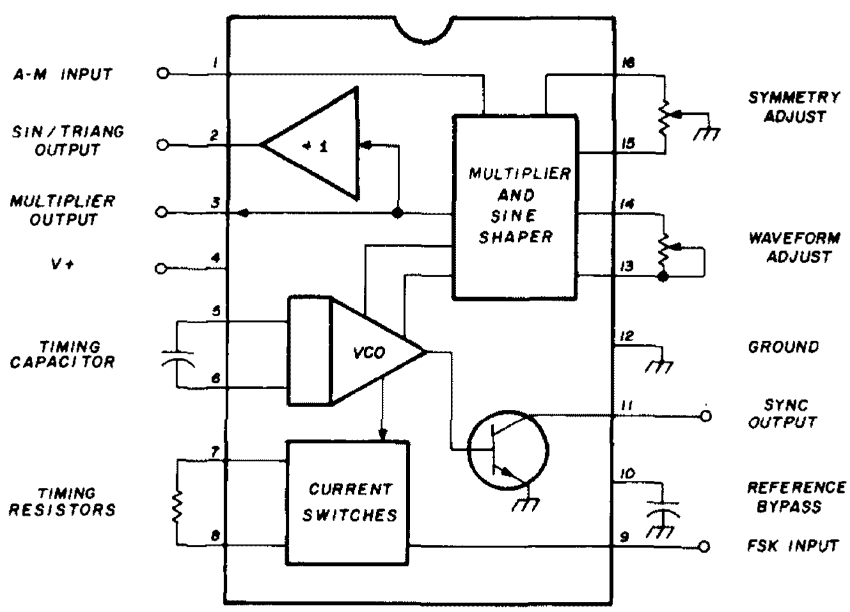


fig. 5. The Exar 2206C function-generator IC.

of 140 dB can be obtained with the use of three crystal filters and three IC gain blocks. The staggering of crystal frequencies and the alternate gain-loss plan provide a high-gain overall bandwidth with low noise content.

Keeping the noise figure down in the i-f section minimizes the requirement for receiver front-end gain. As an aid in this objective the first amplifier of the i-f amplifier can be a field-effect transistor.

In practical vhf/uhf single-conversion operation a good i-f choice would be in the 10- to 12-MHz range. This matches well the standard 10.7-MHz i-f. When using a much higher i-f there may be problems with low-order spurious components, especially if the receiver is to operate over the 25- to 60-MHz range. At lower intermediate frequencies there is an additional cost factor and the 5-MHz limit of present monolithic crystal-filter science.

A simplified schematic of the mixer and first i-f amplifier is shown in fig. 2. The vhf signal from the antenna is applied to a matching input transformer and then to a five-section helical preselector that sets the rf selectivity of the receiver. The output of the preselector is applied to the gate of an fet mixer. Local-oscillator power is injected to the source of the fet.

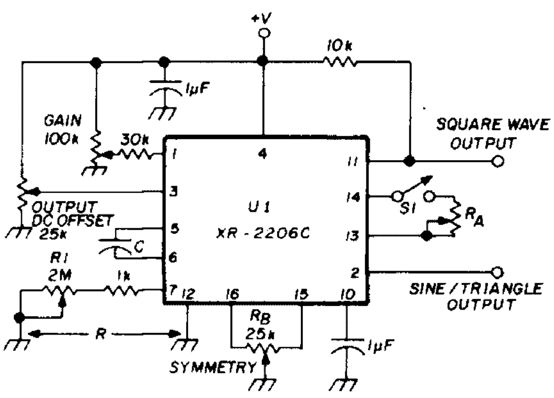


fig. 6. Simple sinusoidal generator using the XR-2206C IC provides sine, triangle or square-wave outputs. Switch S1 must be closed for sine output.

The drain output is linked to the input of the first monolithic crystal filter. An LC network does the appropriate matching to the four-pole filter. A second network matches the output

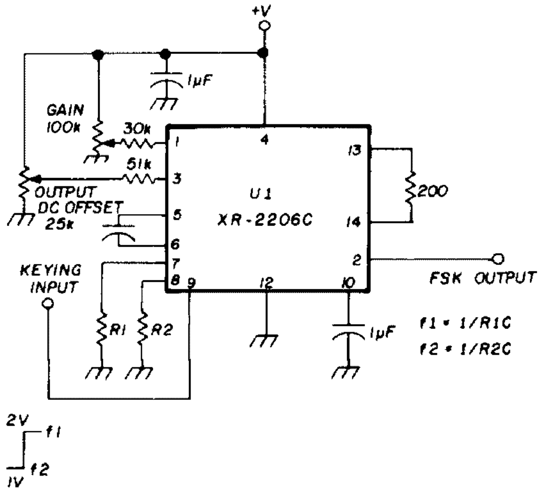


fig. 7. Simple FSK generator using the XR-2206C function-generator IC. Mark (f1) and space (f2) frequencies are given by the equations. The keying input is applied to pin 9.

of the crystal filter to the input of the i-f amplifier. The intermediate frequency is 11.2 MHz and i-f gain is 20 dB.

The first i-f amplifier is followed by a second crystal filter along with the input and output impedance-matching networks, fig. 4. This is followed by an IC i-f amplifier with a gain of 60 dB. The integrated circuit is similar to the RCA CA3014 wideband amplifier IC. No agc system is required. Dynamic signal range, freedom from significant intermodulation distortion components and limiting action down to minimum signal level eliminate this circuit function.

function generator

In the October, 1972, column the Exar 205 and S200 function generators were introduced. These integrated-circuit generators can form a number of basic sine, ramp and pulse waveforms. Generated waveforms can also be ampli-

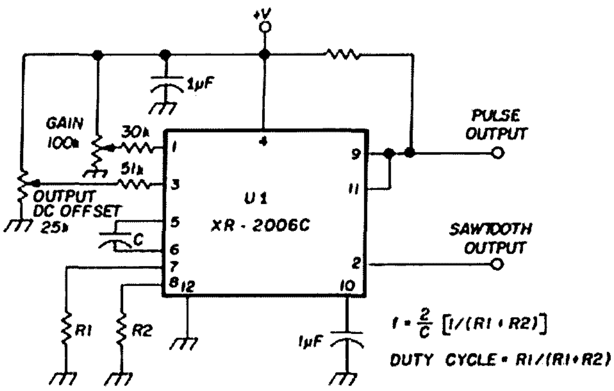


fig. 8. Pulse and sawtooth generator using the XR-2206C function-generator IC.

tude or frequency modulated. Exar now makes a low-cost generator, the Exar-2206C, shown in fig. 5. Similar to the previous function-generator ICs, it comprises four function blocks — voltage-controlled oscillator, multiplier and shaper, buffer amplifier, and current switches. The upper frequency limit is about 1 MHz, less than the previous devices. However, the device has greater stability with temperature change and can be swept over a larger frequency range.

As with the previous devices, only a few external components are needed. A sinewave generator, fig. 6, can be adjusted to provide a pulse, sinusoidal or triangle wave output. The output frequency is determined by the value of the capacitor connected between pins 5 and 6 and the resistance connected between pin 7 and common. The output level is set by the gain control while the offset dc voltage at pin 2 can be adjusted with the 25k offset potentiometer connected to pin 3.

To obtain a sinewave output with minimum harmonic content switch S1 must be closed. Potentiometers R_A and R_B are then adjusted for minimum distortion in the sinewave output. The output frequency is given by

$$f = 1/RC$$

where R is the approximately 2 megohms from pin 7 to ground, and C is connected between pins 5 and 6.

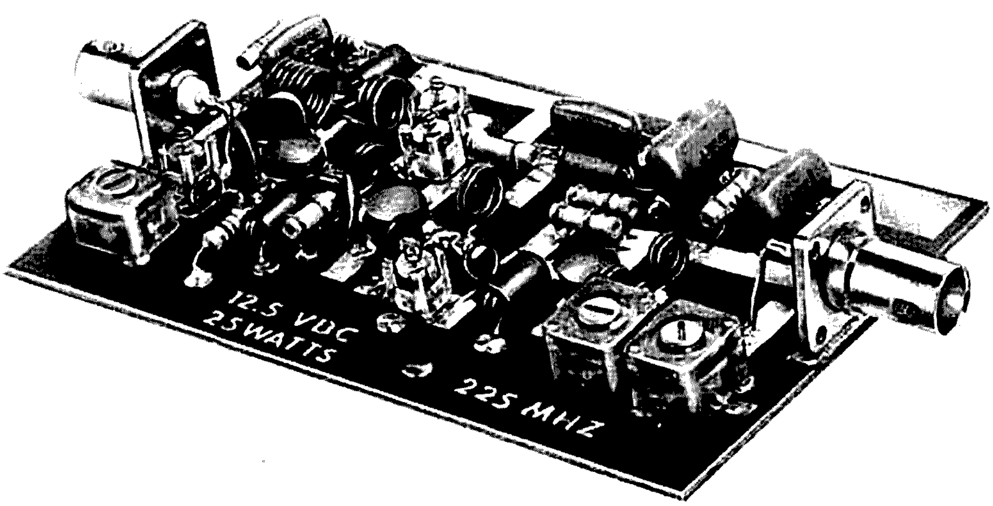


fig. 9. A 25-watt amplifier for 225 MHz which requires 100 mW of drive. Complete circuit of this unit is shown in fig. 10 (photo courtesy Amperex).

A frequency-modulated output can be obtained by applying a modulating wave to either pin 7 or pin 8. Amplitude modulation is obtained by applying the modulating wave to pin 1.

The device can also be used as a simple FSK generator using the circuit of **fig. 7**. In this case the *mark* and *space* frequencies are determined by resistors R1 and R2. The actual FSK keying waveform is applied to pin 9. The mark and space frequencies (f1 and f2) are given by

$$f1 = 1/R1C$$
$$f2 = 1/R2C$$

A circuit for using the 2206C as a pulse and sawtooth generator is shown in **fig. 8**. In this mode pin 9 is shorted to the squarewave output at pin 11. The ramp rise and fall times or pulse duty cycle are regulated with resistors R1 and R2. The pulse width and duty cycle can be adjusted from 1% to 99% by proper selection of values for resistors R1 and R2 as given by

$$f = 2/C\left(\frac{1}{R1 + R2}\right)$$

$$\text{duty cycle} = \frac{R1}{R1 + R2}$$

vhf power amplifier

Amperex has announced a 25-watt, 225-MHz power amplifier module designed for fm circuits, **fig. 9**. This module uses the 2.5-, 8- and 25-watt Amperex semiconductors shown in the schematic diagram of **fig. 10** and features 50-ohm input and output. With a 100-mW input signal, the module will deliver 25-watts output. The circuit is straightforward, using four capacitive dividers for input, output and interstage matching. The collectors are shunt-fed and three decoupling networks avoid self-oscillation. The amplifier can withstand an output mismatch as high as 50:1 without harm.

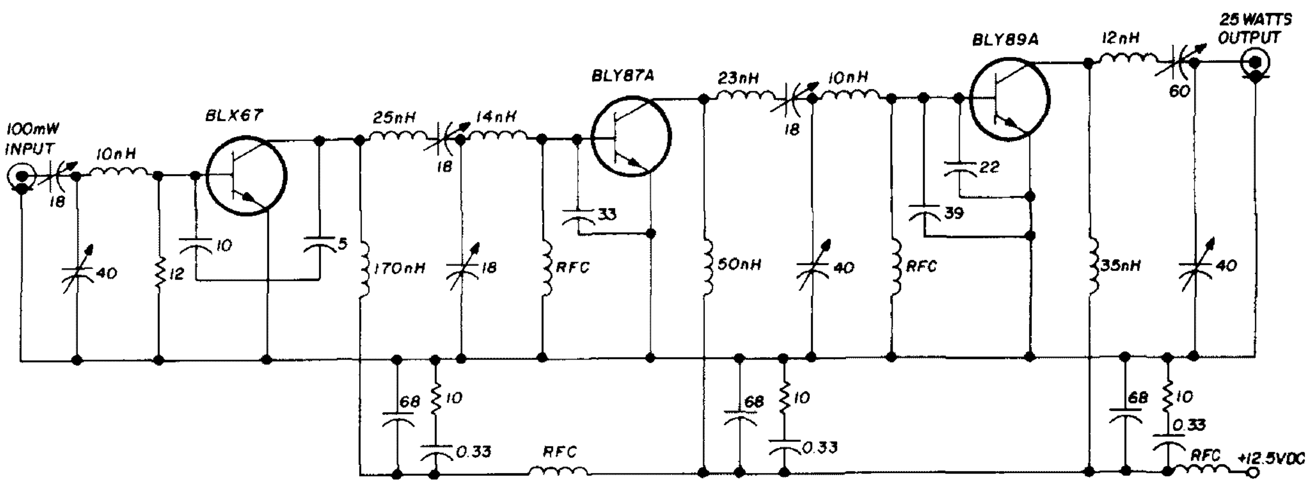


fig. 10. Three-stage solid-state 25-watt power amplifier for 225 MHz (see fig. 9).

metric tapes

Some months ago I mentioned the convenience of a metric tape for calculating antenna and transmission-line lengths. Steel tapes with lengths up to 100 meters can be purchased from Forestry Suppliers, Box 8397, Jackson, Mississippi 39204. Standard lengths of 3, 20, 30, 50 and 100 meters are available. Some models are calibrated in feet and inches as well.

ham radio

the ham notebook

more power from the Standard 826M

Recently I offered to change the final transistor in a Standard 826M for a RTTY enthusiast friend. The output stage had "passed on," not from accidental abuse, but from overwork. Duty cycle or not, 40 minutes of RTTY was a normal day's work, before breakfast!

When I received the unit and instruction manual, I began to study the schematic. Companies such as Standard, that also produce marine and commercial land mobile gear, tune the slugs a little further into the coils and call the set an amateur radio. This is good in many respects. It not only assures that the design is fairly sound; it also lends itself to the plagiarizing of features from one model to another.

The case in point is the power selector system. This feature was originally designed into the marine units by FCC directive. Marine vhf sets must have a power selector system which will select either full power out or slightly less than 1 watt out for inner harbor communication. When this feature was incorporated into the amateur line, it was labeled as a "battery-saver circuit."

The circuitry is similar to the tune/operate switch in older tube and hybrid radios which reduces power output during tuneup. In the Standard, the power selector switch inserts a resistor in the collector line of the final and predriver. Changing the value of this resistor, within reason, makes changing the power output and increasing the duty cycle easy.

My decision on what level of power output to allow was hampered by several obvious and other, not-so-obvious, constraints. First, physical constraints: as the allowed power rises, the value of dropping resistor increases *in size due* to the higher power dissipation. Second, electrical limitations: the wire connecting the dropping resistor to the front panel switch is not suitable for more than several hundred milliamperes.

After some bench and field tests, an output of 2 to 3 watts in the low power position was found to be a satisfactory level. By Ohm's law, decreasing the resistor by a factor of two doubles the current and power output. In the Standard 826M the original resistor value of R005 is 22 ohms. The same value is used in the 806M; in the newer 826MA, the value of R363 was reduced to 15 ohms (consult *individual* instruction manual for value and position).

The resistor will probably be located across the two 9-pin test jacks on the rear of the chassis or on the transmitter board between the chassis wall and output transistor heatsink. A word of caution when working around the final area: take care not to deform the air-wound coils; sometimes they don't bend back. I know from experience!

Simply halving the present value will

produce a 2-watt output. Reducing the value slightly again will bring this to 3 watts. Reducing the resistance below this will increase the heat dissipation of the resistor to too great a value. The replacement resistor should be a Brown Devil or similar type, with a 12 watt, or slightly higher, rating. Mount the replacement in exactly the same fashion as the original unit.

This small modification adds much to the versatility of the 826. First, it provided the longer duty cycle originally sought. Second, some of the new 2-meter bricks need only 2 to 3 watts in to give 40 or more watts out. This means a selectable power of 2, 10 or 40 watts. Third, the power output stability and thermal characteristics of the final device are greatly increased, even in key-down sessions that last 45 minutes.

Although this article has discussed only the Standard line, examination of any 2-meter rig with a low-power switch will produce the same findings. Again, don't shoot for more than 3 watts at the low power setting. It can be done, but not without pulling heavier wires. And watch out for those air-wound coils.

John Pakusich, WB6KVF

open filament pins on power tubes

During a recent operating session, it suddenly appeared that my 3-500Z had died, with an apparent open filament. I replaced it with an elderly 4-400A, kept for emergencies, and let a few choice words out.

Close examination of the 3-500Z was puzzling, since the filament appeared intact. A check with an ohmmeter indicated continuity. Since the 4-400A replacement was working fine, the filament transformer was above suspicion.

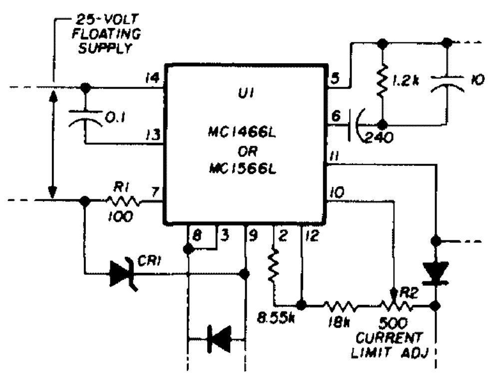
A close look at the filament pins of the 3-500Z, however, lead to the solution. One of the two pins appeared to have lost its solder. The other seemed okay. The bad pin was heated with a 40-watt iron and removed from the heavy filament lead. The filament lead was cleaned and scraped and then the pin was fitted back over it and resoldered. The tube has been successfully restored to full service.

The solder connection was probably not too good to start with, and the heavy starting surge current undoubtedly caused the connection to deteriorate further. Since discovering this malady I have been advised of the same problem with a 4-1000A, and I later saved another 3-500Z from becoming a lamp base. If that expensive tube in your final amplifier quits, be sure to check the solder connections to the filament pins before consigning it to the trash can.

Bob Locher, W9KNI

short circuit

In the circuit for the regulated solid-state high-voltage power supply described in the January, 1975, issue (**fig. 1**, page 42), the 18k resistor connected between pin 2 and the 8.55k resistor should be connected to pin 12; the 8.55k resistor should be connected between pins 2 and 12 as shown below. Also, the 0.33



μF capacitors connected from the secondary of transformer T3 to ground should be rated at 600 volts.

new products

Heath SB-104 series

The new SB-104 series of equipment from Heathkit is one of the most advanced designs in amateur radio today and includes such features as digital readout, all solid-state circuitry, and broadband rf design. Key units of equipment in the new Heath line are the SB-104 five-band transceiver, SB-644 remote vfo, SB-634 station console, SB-614 station monitor and SB-230 conduction-cooled kilowatt linear.

transceiver

The new SB-104 ssb and CW transceiver is completely solid-state from front end to rf output and runs a cool 100 watts output. The four final transistors are completely protected against high vswr and thermal runaway so there is practically no danger of damaging them in normal operation. The Heath Company is so sure of the design, in fact, that they have placed a one-year warranty on the rf output board and final transistors. Since all of the rf circuits of the SB-104 are broadband, you can move instantly from one end of the band to the other, or from band to band; there is no need to adjust any preselector, drive, load or turn controls — just choose the band you want, dial in your operating frequency, and go on the air.

The digital dial of the SB-104 provides direct 6-digit readout of your operating frequency with 100-Hz resolution on all bands. And, unlike some digital readout systems that actually read only the vfo frequency (and interpret the output frequency), the digital dial in the SB-104 accounts for the vfo, bfo *and* high-frequency oscillator signals. With this arrangement there is no need for a frequency calibrator.

The large spinner dial provides about 30 kHz per revolution, a rate that seems ideal for all types of operating habits. Of course, the 100-Hz accuracy of the digital dial means that when you want to be on a certain frequency, that's where you'll be. If you only need 1-kHz accuracy, the last, 100-Hz digit can be turned off with a front-panel switch.

Although the final amplifier of the SB-104 is rated at 100 watts *output* (ssb or CW), for QRP operation the output can be instantly switched to one-watt with a front-panel pushbutton. The carrier and unwanted sideband are suppressed 55 dB, harmonic radiation is down 45 dB, spurious radiation within ±3 MHz of the carrier is down 50 dB or more, and third-order intermodulation distortion is 30 dB down from two-tone output.

The broadband receiver has been designed for minimum cross-modulation and intermodulation distortion products — important aspects on today's crowded amateur bands. Adjacent signal overload is negligible with the SB-104 yet sensitivity on all bands is 1 μV or better for 10 dB signal-plus-noise-to-noise ratio. Selectivity is 2.1 kHz minimum at 6 dB down and 5 kHz maximum at 60 dB down (2:1 shape factor). With the accessory 400-Hz CW filter, selectivity is 400 Hz at 6 dB down and 2 kHz maximum, at 60 dB down. Audio output is rated at 2.5 watts into 4 ohms at less than 10% THD (1 μV input signal provides 0.5 watt audio output). Agc is switch select-

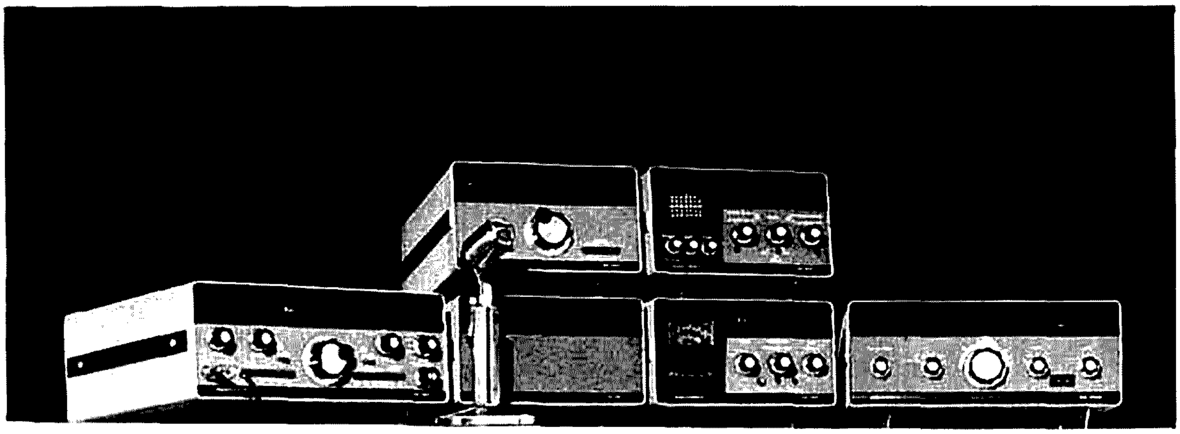
More Details? CHECK—OFF Page 110

able for release times of 100 μs or 1 millisecond; attack time is less than 1 millisecond.

Front panel controls include agc (fast, slow, off) audio and rf gain, main tuning, mic/CW level, vox gain, vox delay and bandswitch. Pushbutton switches are used to control vox, noise blanking, mode, tune, high/low power, and metering. On the rear panel are controls for vox anti-trip and sidetone level. Rear panel connections are provided for phone patch, auxiliary audio output, speaker, key, alc, external vfo, i-f output and separate receive and transmit antennas.

The more than 2800 parts used in the SB-104 transceiver are mounted on 15 separate printed-circuit boards for easy assembly and test. Eleven of the boards plug-in and seven of these may be extended out of the chassis for adjustment or troubleshooting while the transceiver is operating. Two large wiring harnesses eliminate 95% of the point-to-point wiring in the SB-104. Total construction time is about 50 hours. Alignment of the completed transceiver is fast and simple, requiring only a dummy load, microphone and vtvm.

The SB-104, priced at $699.95, is designed for use with a 13.8-volt dc power supply so it is a natural for mobile operation. Current drain is 2 amps on receive and 20 amps on transmit (3 amps when switched to low power). The HP-1144 fixed-station power supply ($89.95) provides the necessary voltages from 120/240 Vac lines. Other accessories for the SB-104 include the SB-604 station speaker ($29.95) SBA-104-1 noise blanker ($24.95), SBA-104-2 mobile mount ($34.95), and SBA104-3 400-Hz CW filter ($34.95).



remote vfo

The SB-644 remote vfo ($119.95) was designed specifically for the SB-104 transceiver and provides serious DXers with complete split transmit/receive capability. The transceiver vfo can be at one end of the band while the remote vfo is at the other. Furthermore, the system is designed for transceive operation on the remote vfo *or* internal transceiver vfo, transmit on the SB-104 and receive on the remote, or receive on the SB-104 vfo and transmit on the remote. There are also provisions for two crystals in the SB-644 for fixed-frequency control.

Although the linear dial on the front panel of the SB-644 remote vfo places

you in the right ballpark, actual frequency readout is provided on the transceiver's digital display. The display automatically changes to the correct frequency as you switch from receive to transmit so there's never any doubt as to the frequency you are operating on.

Four front-panel pushbutton switches on the SB-644 control all transceive, transmit and receive modes on both the transceiver and the remote vfo. No switching is necessary at the transceiver — just push a button to turn on the vfo, and push another to select the vfo or crystal. Status lamps behind the window indicate whether the frequency is controlled by the transceiver vfo, remote vfo or crystal.

The vfo circuitry in the SB-644 is the same as that used in the new SB-104, and thanks to the true digital readout used in the transceiver, concern about vfo dial linearity is a thing of the past. The vfo kit is assembled on two circuit boards and only two simple adjustments are required for alignment. Frequency drift is less than 100 Hz per hour after a 30-minute warmup; dial backlash is 100 Hz maximum.

station console

The new SB-634 station console ($179.95) is actually five station accessories in one: a 24-hour digital clock, ten-minute ID timer; phone patch, rf wattmeter and swr bridge. The digital clock indicates hours, minutes and seconds with digits large enough that they can be read from across the room. The clock runs continuously as long as the console is plugged in, completely independent of the other functions.

The ten-minute ID timer uses three digits to indicate minutes and seconds up to 9:59. At the ten-minute mark the timer automatically recycles and provides a visual alarm or both visual and audible alarms (selectable from the front panel). If you elect to identify before the 10-minute mark, as you

More Details? CHECK–OFF Page 110

should, the ID timer can be reset by a pushbutton on the front panel. Accuracy of both the 24-clock and ten-minute timer is determined by the accuracy of the power-line frequency.

The hybrid phone patch which is built into the station console allows either manual operation or voice control without switching connections. Isolation between transmit and receive circuits is at least 30 dB and can be adjusted with a rear-panel control. When you are running a phone patch, the meter can be used to indicate VU, and transmitter and receiver gain can be set with separate front-panel controls.

The rf power meter covers the frequency range from 1.8 to 30 MHz in two ranges, 200 or 2000 watts full scale, with accuracy of ±10%. If you wish to measure swr, simply press a button on the front panel. Swr sensitivity, which is less than 10 watts, is adjustable.

station monitor

The Heath SB-614 station monitor ($139.95) will display transmitted ssb, CW, a-m (trapezoid) and RTTY (cross) signals up to one kilowatt from 80 through 6 meters. The flat-face CRT uses push-pull drive for a keystone-free, sharp, clean trace that can be used to diagnose a wide variety of operating problems: non-linearity, insufficient or excess drive, poor carrier or sideband suppression, regeneration, parasitics and key clicks. The operating manual includes 40 CRT display illustrations and explanations.

The station monitor has all standard scope control functions with a recurrent, automatic sync-type sweep generator that is adjustable in three ranges from 10 Hz to 10 kHz. For limited test applications the SB-614 can also be used as a normal scope with 10 kHz to 50 kHz bandwidth, good sync and high input sensitivity (60 mV rms for ¼-inch vertical deflection). A rear panel 10:1 attenuator provides extra operating convenience.

conduction-cooled linear

The new SB-230 1-kilowatt conduction-cooled linear uses a rugged Eimac 8873 triode in a proven grounded-grid circuit to deliver up to 1200 watts PEP input on single sideband and 1000 watts on CW with less than 100 watts drive. The linear is also rated at 400 watts input for slow-scan television or RTTY. Third-order intermodulation distortion is −30 dB or more. And since the SB-230 is conduction cooled, there is no need for a noisy blower – the massive heatsink on the rear of the unit takes care of all the cooling requirements.

The cabinet of the SB-230 features microswitch interlocks on both the top and bottom to shut down the primary power when either of the covers is removed. The temperature of the heatsink is monitored so if the temperature rises too high a thermal circuit breaker opens and the amplifier shuts down. To allow the 8873 sufficient time to warmup when it is first turned on, a delay circuit is built in. When warmup is completed the Delay light goes out. Lights are also provided to indicate when the exciter is running straight through and when the amplifier has been shut down because of high heatsink temperature.

Bandswitching of the linear is accomplished with a single knob and the load and tune controls are clearly marked so you can return to a favorite operating frequency by simply noting the control positions. The back-lighted meter is used to indicate relative rf power, plate current, grid current or high voltage. Relative power sensitivity is adjustable from the front panel. Price, including power supply, is $319.95.

For more information on the exciting new Heath SB-line, the first complete high-frequency amateur communications system to be offered in some time, write to the Heath Company, Benton Harbor, Michigan 49022, or use *check-off* on page 110.

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$1.00

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focus on communications technology . . .

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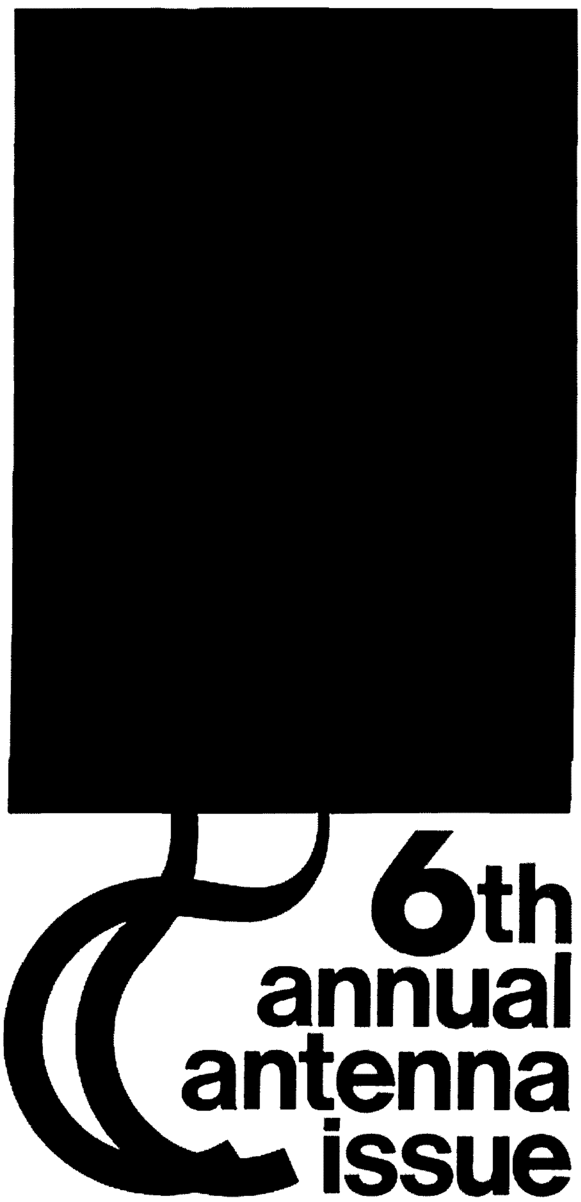
MAY 1975



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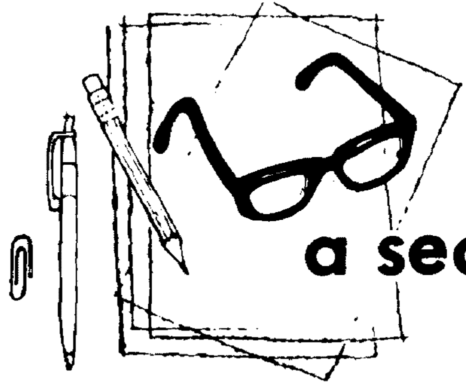


a second look by jim fisk

Power measurement of amateur transmitters is a serious topic of discussion at the FCC these days. Although the amateur restructuring proposal, Docket 20282, has proposed that the maximum transmitter *output* power shall not exceed 2000 watts peak envelope power, a footnote points out that this is only one proposal under consideration by the FCC. The Commission is also looking at alternatives such as PEP input, average power input, ratios of peak to average power output, limitations on the dissipation ratings of final power amplifier devices or even a combination of these. Prose Walker, W4BW, Chief of the Amateur and Citizens' Division, is soliciting suggestions for solutions to this problem, and doesn't want to wait for the Docket comments to hear your ideas. A note to him at 2025 M Street, NW, Washington, DC 20554, with your recommendation, would be appreciated.

Although some amateurs may feel that it makes little difference whether the regulations specify input or output power, at frequencies above 50 MHz or so, where operating efficiencies are much lower, it is extremely important. On the high-frequency amateur bands typical transmitter efficiency is on the order of 70 per cent, but at vhf and uhf efficiency can drop to 50 per cent or even less. That 20 per cent difference can make a considerable difference to the vhfer who is trying to extend his communications range. And, with the rf wattmeters which are currently available, there seems to be little reason why the amateur power limitations shouldn't be expressed in terms of output power.

Until relatively recently, the accurate measurement of high-frequency rf power required very expensive instruments to which few amateurs had access: matched dummy loads, calorimeters and other thermal devices which could be calibrated with known dc voltages and currents. Furthermore, the difficulty of rf power measurements increased with both frequency and power level. Accurate measurement of 500 watts output, for example, was many times more difficult than the measurement of 50 watts output, even on the hf bands. These measurements were further complicated at vhf and uhf, and accurate measurement of the output of *high-power* vhf transmitters was virtually impossible.

The introduction of toroidal-type directional wattmeters about fifteen years ago solved the problem of high-frequency *average* power measurement (subsequent design improvements have moved the frequency range up to 250 MHz or so), but the difficulty of measuring peak envelope power still remains. At least one coupler (the Collins 302C) added a capacitor to the circuit to make it more of a peak-reading device, but it still reads only about 65 per cent of the actual peaks. Most other types of wattmeters, including absorption types and the popular Bird model 43, theoretically read 40.5 per cent of the actual peak power of a two-frequency signal. The Bird model 4311 reads peak power directly, but it's priced out of the range of most amateurs. However, it should be fairly simple to design a peak-reading amplifier for use with average-reading rf wattmeters that would provide direct measurements of peak envelope power.

Another possibility is the use of a single-tone. Single-tone modulation of a ssb transmitter results in peak envelope power output which equals average power and is indicated correctly on average-reading power rf power meters. However, this technique is not nearly as tidy as a peak-reading rf wattmeter which is always in the line.

Jim Fisk, W1DTY
editor-in-chief

presstop

BICENTENNIAL CALLSIGN SCHEME has been released, and as expected, all FCC-licensed amateurs can have some fun with it in 1976 if they're so inclined. Stateside alternate prefixes are straight forward, with WA becoming AA, W changing to AC, and repeaters getting an optional AF.

Outside The Contiguous U.S. is where the fun really begins, however, with Samoa becoming AH3 and an Alaskan Novice an AL1. The partial scheme appears below. Don't rush to cut it out because we'll include a larger scale chart later this year.

Use Of The Alternate Prefixes is entirely optional and no paperwork will be required. Just remember they don't go into effect until 0500Z January 1, 1976, and are good until January 1, 1977.

WA1-WAØ	AA1-AAØ	KB6	AG2	KP4	AJ4
WB1-WBØ	AB1-ABØ	KC4	AL4	KP6	AIØ
W1-WØ	AC1-ACØ	KG6	AG6	KS4	AH4
K1-KØ	AD1-ADØ	KH6	AH6	KS6	AH3
WD1-WDØ	AE1-AEØ	KJ6	AJ7	KV4	AJ3
WR1-WRØ	AF1-AFØ	KL7	AL7	KW6	AG7
WN1-WNØ	AG1-AGØ	KM6	AH7		

CLASS-E CB DECISION DEFERRED again, will be reconsidered "later this year" as a result of March FCC meeting. Commissioners decided that Class-E CB issue could not be properly decided without taking both amateur and Class-D CB "restructuring" dockets as well as the new automatic transmitter ID docket under consideration at the same time, and Reply Comments on the latest of these aren't due until July 16. Canada's objections to 220-MHz CB were also a factor.

COMMENTS ON DOCKET 20282 are arriving in a steady stream at the FCC, and several hundred have been received so far. Though no tabulation is being kept, most from Extras complain about loss of the exclusive phone bands and those from Generals and Techs bemoan their many losses. Very few flatly reject the proposal.

AMSAT EXPERIMENTER'S CONFERENCE in late March covered lots of ground, and tentative plans are for a "Phase 3" spacecraft on an elliptical orbit for launch in 1978. Like its predecessors, the new satellite would be a truly international project: design work is to be German, construction Canadian and ground support Australian.

Tech Class Licensees can use OSCAR 7, despite report to the contrary in east coast club paper — specific permission was a part of the FCC license issued to the satellite.

AMSAT Dues Will Double July 1 from $5 to $10 a year, so it would be wise to renew now for an extended period and save.

OSCAR 6 Telemetry supports the expectation that AMSAT's long lived "bird" will continue to be operational at least through next fall. Battery temperature, considered to be the most critical parameter, is dropping after the expected February peak.

6-METER BAND THREATENED by proposal to add another VHF TV channel. In its comments filed in response to Docket 20264, which pertains to radio call-box systems and their use in the 72-76 MHz band, a prominent Dallas consulting engineering firm has proposed that the public interest would be best served by moving channels 2, 3 and 4 down 2 MHz and putting a new TV channel, dedicated to non-commercial educational use, in the resulting 70-76 MHz band. Present users of the 72-76 MHz slot would move elsewhere, but the other 2 MHz would come right out of the amateur 50-54 MHz assignment!

Whether This Proposal will receive serious consideration by the Commissioners remains to be seen, but this threat — very similar to those of EMS on the high end of 420-450, HIRAN and 220-MHz Class-E CB, shows how others look at the present amateur assignments.

large vertical antenna for 160 and 80 meters

A design approach for dealing with problems of erecting an efficient radiator to compete on the two lower bands

Harry R. Hyder, W7IV, Scottsdale, Arizona 85253

The decrease in sunspot activity has caused a renewed interest in low-frequency DX operation. On the 80- and 160-meter bands beams are impractical, and even dipoles must be unreasonably high to get the low-angle radiation necessary for DX. That leaves vertical antennas. Although omnidirectional, they are very good low-angle radiators. In theory, a short vertical radiates as well as a tall one; but in practice, the low radiation resistance of a short vertical, compared with ground and other resistances, makes its overall efficiency low.

My friend Liscum Diven, W7IR, decided to erect a good-sized vertical hoping to increase his DX contest scores on 80 and 160. He is blessed with enough real estate to make this practical, considering the area required for guying and a good ground system. While all amateurs are not so fortunate, the design procedures described in this article are applicable to more modest antennas — or larger ones.

height considerations

The first thing to be decided, of course, was the height. This factor is always a compromise between desired performance and cost. There is no good reason why a vertical, or any other antenna, has to be self resonant. An antenna will radiate all the power it can absorb. When an antenna is an odd multiple of a quarter-wavelength long the feed-point impedance is conveniently low for coax transmission lines, and the reactance is zero; but these are the only advantages to resonant antennas: they are easier to feed.

As for the vertical, the practical efficiency increases until a height of about 0.6 wavelength is reached. At greater heights the vertical radiation angle rises rapidly, making the antenna less desirable for DX.

Since W7IR's antenna was to be used on both 80 and 160, the height could not exceed 0.6x80 meters, or 157 feet (48m). This was a little too high for W7IR's tastes. Top-loading to reduce the physical height while maintaining the electrical height was considered but rejected because of mechanical difficulties.

Even if you're not particularly interested in erecting a vertical antenna, the following piece is well worth reading for a firm grasp on some of the physical aspects of all antennas. The part on characteristic impedance, which is treated in terms of an antenna as a transmission line, should help to dispel some of the misconceptions on antenna theory that we hear on the amateur bands. The graphs of resistance as a function of antenna height-to-diameter ratio, which were taken from reference 1, should provide a convenient design aid for those seriously contemplating the construction of a large vertical antenna. **Editor**

It was finally decided to make the vertical 91 feet (28m) high, consisting of 70 feet (21m) of aluminum lattice tower sections, surmounted by a 21-foot (6.4m) whip that happened to be

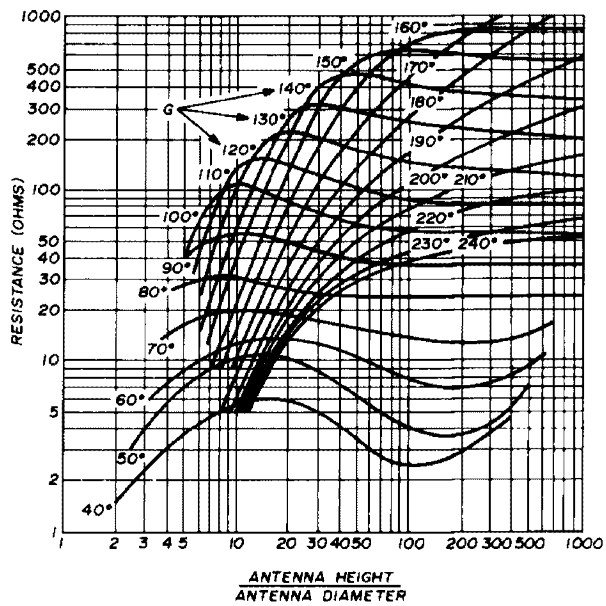


fig. 1. Resistance as a function of antenna height-to-diameter ratio.

available. This configuration would make the electrical height of the antenna 0.347 wavelength on 80 and 0.166 wavelength on 160 meters.

It was decided to guy the antenna at two levels. The base insulator was to be a cluster of heavy-duty ceramic pillars. Since the dead weight of the antenna would be only about 100 pounds (37kg), and ceramic is very strong in compression, this type of base was quite practical.

The guys were to be 1/8 inch (3mm) diameter steel cable, with 6-foot (1.8m) sections of 1/2-inch (13mm) polypropylene rope to insulate the guys from the tower, and small egg insulators to break up the guys. The guy anchors would be 5-foot (1.5m) long earth augers.

matching system characteristics

A lot of thought was given to the matching network. It was decided not to use the usual cut and try method; instead, the network would be engineered. The network was built before the antenna was erected, and required only minor adjustment when installed.

The following characteristics were desired in the network:

1. Obviously it must match the resistance and tune out the reactance. The antenna would be highly reactive on both bands since it would not be resonant on either.

2. It should use as few elements as possible and should be easily switched between bands.

3. It should have a permanent dc path to ground on both bands for lightning protection.

The design achieved all of these objectives.

First to be determined was, "what were we matching?" Any antenna is really a transmission line. Its termination is its own losses to space; but since this is a poor termination, the vswr is very high, as a graph of the voltage and current shows. As with all transmission lines, an antenna has a characteristic impedance. This is *not* the sending-end impedance. What is seen at the sending end is the antenna's losses to space, which are distributed along its length, transformed by the characteristics of the transmission line to a single resistance value called the base radiation resistance of the antenna. This resistance, when multiplied by the square of the current measured at that point, tells how much power is actually being radiated.

The characteristic impedance of the antenna is a function of the length-to-diameter ratio. A thin wire will have a characteristic impedance of 600 ohms or more; a lattice tower might have a characteristic impedance of 200 ohms or less. The lower the characteristic impedance, the less will be the impedance excursions with frequency at the sending end, which is true of all trans-

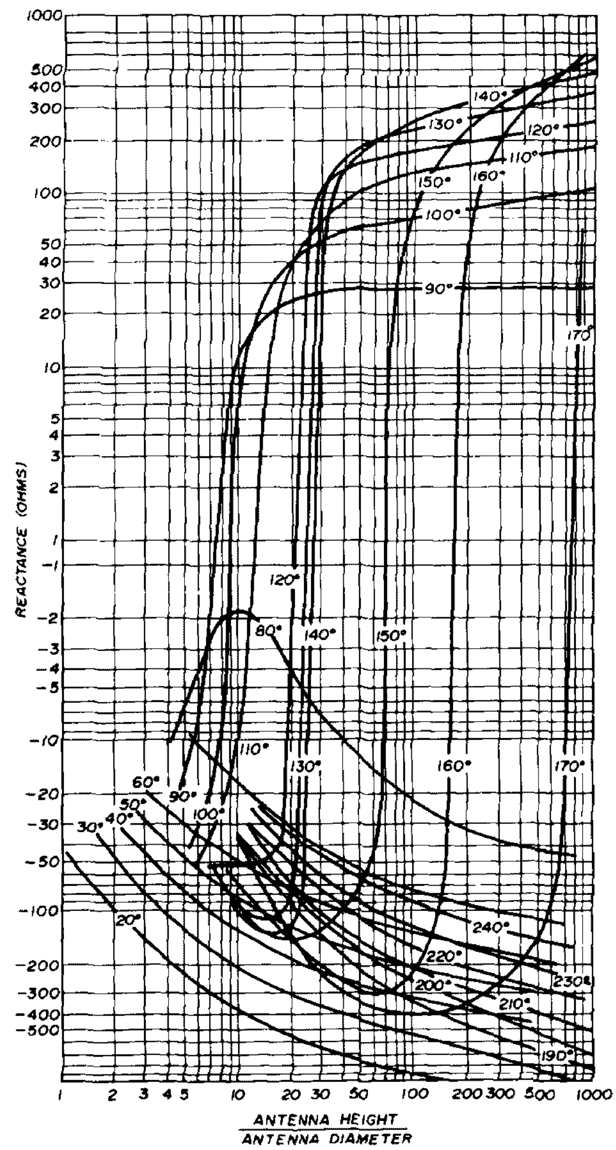
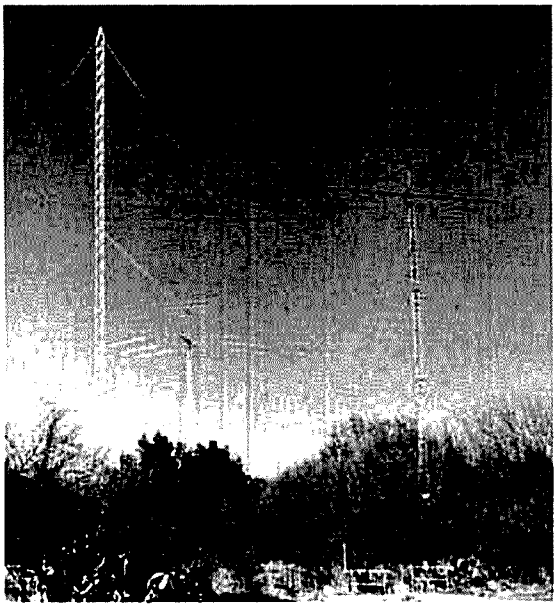


fig. 2. Reactance as a function of antenna height-to-diameter ratio.

mission lines. This means greater bandwidth.

The accurate calculation of the resistance and reactance of an antenna of irregular shape is impractical, but graphs are available that give fairly realistic

values. These graphs, reproduced in figs. 1 and 2 and described in reference 1, were generated from vhf scale models and give resistance and reactance values for solid cylindrical antennas of different height-to-diameter ratios over a perfect ground, with the height of the antenna in electrical degrees as the parameter.



Antenna farm at W7IR includes beams for 20 and 15 meters, right, beams for 40 and 10 meters, center, and 91-foot vertical for 80 and 160 meters, left.

design assumptions

We didn't know how a lattice tower of triangular cross-section related to a solid cylinder, but we guessed that it might be something like the diameter of a circle inscribed within the 11-inches (28cm) on-a-side triangular mast or 6 inches (15cm). We also thought that the top 21 feet (6.4m) of the antenna, which was a small-diameter whip, would reduce this dimension further. Since it was a nice number to work with, our final guesstimate was a height-to-diameter ratio of 200, which proved to be a good choice. Because W7IR's antenna would be 0.166 wavelength on 1.8 MHz and 0.347 wavelength on 3.75 MHz, the electrical heights on those bands would be 59 and 125 degrees respectively.

The graphs, as closely as they could be read, yielded the following R and X values for the two bands:

frequency	resistance, R (ohms)	reactance, X (ohms)
1.8 MHz	7	-160
3.75 MHz	180	±240

We further assumed that the ground system, fair but far from ideal, might represent a loss of 3 dB at 1.8 MHz. Adding 7 ohms, a loss resistance equal to the 1.8-MHz radiation resistance, might compensate for ground-system losses. The effective resistance values thus would be 14 and 187 ohms, respectively.

The antenna would require a series inductive reactance of 160 ohms to tune it to 1.8 MHz, but a shunt reactance was desired to provide a direct path to ground for lightning protection. So the series R and X from the graphs were translated to their equivalent parallel values:

$$Q = \frac{X_s}{R_s}$$

$$R_p = R_s\,(Q^2 + 1)$$

$$X_p = \frac{R_p}{Q}$$

This gave the following values:

frequency	resistance, R (ohms)	reactance, X (ohms)
1.8 MHz	1820	-160
3.75 MHz	495	+386

Thus a shunt reactance of +160 ohms (14.1 μH) would tune the antenna. The parallel resistance value of 1820 ohms could most easily be matched by using

the loading coil as an autotransformer, connecting the 50-ohm transmission line to a tap on the coil. The fraction of coil turns across which the line should be connected was:

$$\sqrt{\frac{50}{1820}} = 0.166 = 16.6\%$$

neglecting leakage inductance. This took care of the 160-meter band.

A shunt capacitor would be required to tune the 80-meter band. This circuit would not furnish a direct path to ground, so it was decided to see what would happen if the 14.1-μH 160-meter loading coil were left in place on 80 meters. At 3.75 MHz this coil would present a reactance of +332 ohms. This reactance in parallel with the +386 ohms reactance of the antenna at 3.75 MHz gave a net reactance of +179 ohms, the resistance being unchanged. When this combination was translated back to series form, the reactance was R = 57.5 ohms and X = +159 ohms, a convenient value since it meant that on 80, with the 160-meter loading coil in place, the transmission line could be connected directly to the base of the antenna through a capacitive reactance of 159 ohms (267 pF) with negligible mismatch. The two-band matching network was thus very simple, consisting of a single inductor and a single capacitor with a spdt relay to switch bands. The network is shown in **fig. 3**.

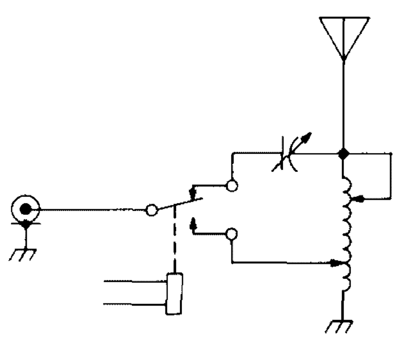


fig. 3. Antenna tuning network for the 2-band vertical with switching relay for changing bands.

W7IR works both phone and CW on 80, so the bandwidth was calculated by obtaining resistance and reactance values for the antenna and network components at 3.5 MHz and 4 MHz and plotting the results on a Smith chart. This data showed that the vswr would be less than 2:1 over the entire band.

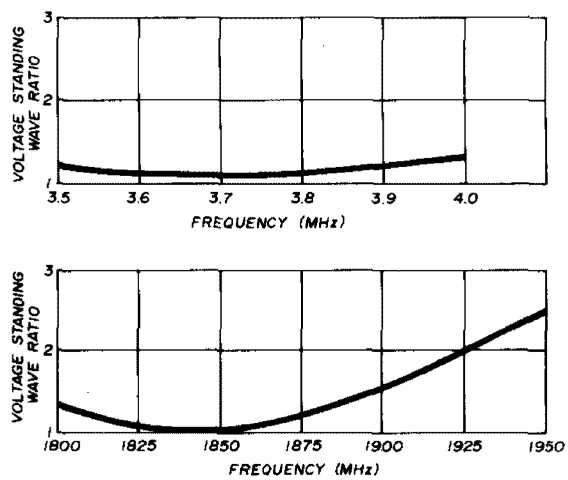


fig. 4. Vswr performance of the large vertical antenna on 80 and 160 meters.

construction and tuneup

The network was built into an old breadbox; components were from an old rig. The coil had a Q of over 200 on both bands, so its loss was negligible. The antenna was erected without mishap. A sign-erection truck was hired for the occasion, and a few local hams held guy wires.*

The radials were put in with the aid of a *mole,* which is a tool about the size of a power lawnmower designed for digging shallow trenches for the pipes of underground sprinkler systems. Sixteen

*Also try your local phone company or a tree-trimming service. They have mobile cherry pickers with operators that can often be hired for a fairly reasonable fee. **Editor**

radials about 125 feet (38m) long were installed. More radials would have been desirable and probably will be installed later. But it was only a week before the start of the DX contest, and even with the mole it was hard work. Then came the tune up. Would the network act as predicted? It had already been built.

Author W7IV supporting W7IR's antenna. Concrete base is 18 inches (45.7cm) by 1 foot (30.5cm). Tuning network and ceramic insulators are shown.

It was our intention first to measure the resistance and reactance of the antenna, as a check on the curves and to get an approximation of the actual ground resistance. This would be done by subtracting the 7 ohms radiation resistance at 1.8 MHz taken from the graph from the measured total value. This turned out to be impossible for an unforseen reason. A local broadcast station produced a signal of *20 volts* at the base of the antenna, making use of an rf bridge impossible. The BC station would not shut down for us so the network was tuned, first on 160, by energizing the network and antenna with enough transmitter power to overcome the BC signal, then adjusting the inductance for maximum voltage across the coil as read by an rf vtvm. Then the transmission line tap on the coil was selected for minimum vswr. Next, power at 3.75 MHz was fed to the network, and the series capacitor adjusted to null the vswr.

The final values came out extremely close to the calculated values. The 160-meter loading coil turned out to be 14 μH (right on the nose) and the 80-meter series capacitor was 320 pF rather than the 267 pF calculated. The transmission line tap on the coil was at 24% rather than 17% of the turns.

Curves of vswr *versus* frequency are shown in fig. 4. The bandwidth is somewhat greater than calculated. This is not necessarily good, however, since it probably means that the ground losses are higher than assumed.

conclusion

As this is being written, the DX contest has just ended. Contest scores, of course, are highly dependent on conditions and the number of stations and countries participating. But W7IR reports that the overall performance on both bands was far superior to that of the high inverted-vees previously used.

On 160, although activity was sparse and noise levels high, W7IR worked every DX station he heard. On 80, W7IR felt that, for the first time, he had an antenna that really put him in a good competitive position. The antenna gave him the feeling that he was really getting through — every serious DXer knows that's what really counts.

references

1. E.A. Laport, *Radio Antenna Engineering*, McGraw-Hill, New York, 1952.
2. R.E. Leo, W7LR, "Vertical Antenna Ground Systems," *ham radio*, May, 1974, page 30.

ham radio

graphical design method for log-periodic antennas

A no-math approach to the log periodic design problem

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Anyone who has designed or studied log-periodic antennas is aware of the math involved due to the many variables that enter into the design problem. References 1-5 contain several pages of formulas and refer the reader to four or five nomographs, log tables, etc. Probably this is one of the reasons the log-periodic antenna has been neglected by the amateur fraternity. Furthermore, little information has been published on the design of hf L-P antennas in amateur publications.

When I retired in 1970 I decided to make a study of high-frequency log-periodic antennas. The original antenna in use here has only 7 elements, is limited to 20 and 15 meters, is less than 40 feet (12.2m) long, and is pointed south. Over the past three years it has averaged 8-10 dB gain compared with a 20-meter dipole at the same height. The results obtained from this beam prompted a second, larger log-periodic for 20, 10 and 15 meters having a boom length of 70 feet (21.3m). Three log-periodics were erected and tested during 1970, and as of this *writing* 17 have been put up and tested.

This article presents a graphic design approach for log-periodic antennas that eliminates the work associated with the math involved. Four designs are presented first, each having essentially the same boom length and apex angle, but with different numbers of elements. Each should provide approximately 10-dB gain referenced to a dipole at the same location and height above ground.

Additionally, two more log-periodic designs are shown, one with a 54 foot (16.5m) boom that gives about 8-dB forward gain and one with a 100-foot (30.5m) boom. (Both boom lengths are nominal.) This last design is presented for those with enough real estate to accommodate it and the nerve to hang such a monster in the air. It will give 12-dB forward gain (referenced to a dipole) if properly designed and assembled and suspended at least 40 feet (12.2m) above ground. All designs cover 14 to 30 MHz.

the log-periodic antenna

For those readers not acquaintec with the log periodic, it is a broadband multi-element, unidirectional, end-fire array capable of 8- to 14-dB forwarc gain. The front-to-back ratio is usuall) 10 to 14 dB with side attenuation tc about 25 dB. The forward lobe of the

log-periodics tested here generally runs about 90 degrees in the H plane. Its vertical angle of radiation, or take-off angle, can be controlled fairly well by height above ground. The swr at the feedpoint remains relatively constant over the frequency range for which the antenna is designed, generally not exceeding 2:1 with 1.5:1 as typical; usually varying between 1.1:1 and 1.5:1.

A bandwidth of 10:1 is normal for fixed commercial log-periodics designed to cover frequencies between 3 to 30 MHz; however, they are quite long, usually 250-800 feet (76-244m), depending on gain required and beamwidth. Some of the commercial log-periodics having a limited bandwidth of 4-30 MHz are 150-350 feet (46-91m) long, and for 6-30 MHz, 100-250 feet (30-76m) long. The commercial rotary types generally have a boom length of 40 to 74 feet (12-22.5m). Some of these are used at MARS stations.

For amateur applications, a fixed log-periodic wire beam can be limited to a bandwidth slightly more than 2:1, covering 7-14.5 MHz for 40 and 20 meters or 14-30 MHz for 20, 15 and 10 meters without being excessively long. By limiting bandwidth still more, say 14 to 22 MHz, a log-periodic can be erected in a space 40x40 feet (12x12m) giving an 8- to 10-dB gain on 20 and 15 meters.

Most of the log-periodics used here have been of the horizontal dipole configuration and have been tested on 40, 20, 15 and 10 meters. One of the 20-, 15-, 10-meter log-periodics was also tested for a few weeks in the vertical plane. Three of the vertical monopole configurations using a ground plane or counterpoise have been tested on 40 and 80 meters. More recently two of the trapezoidal type, one the sawtooth structure and the other the zig-zag, have been tested on 20 meters. Some of these log-periodic antennas have been described in amateur publications[6-10] with complete dimensions and assembly details.

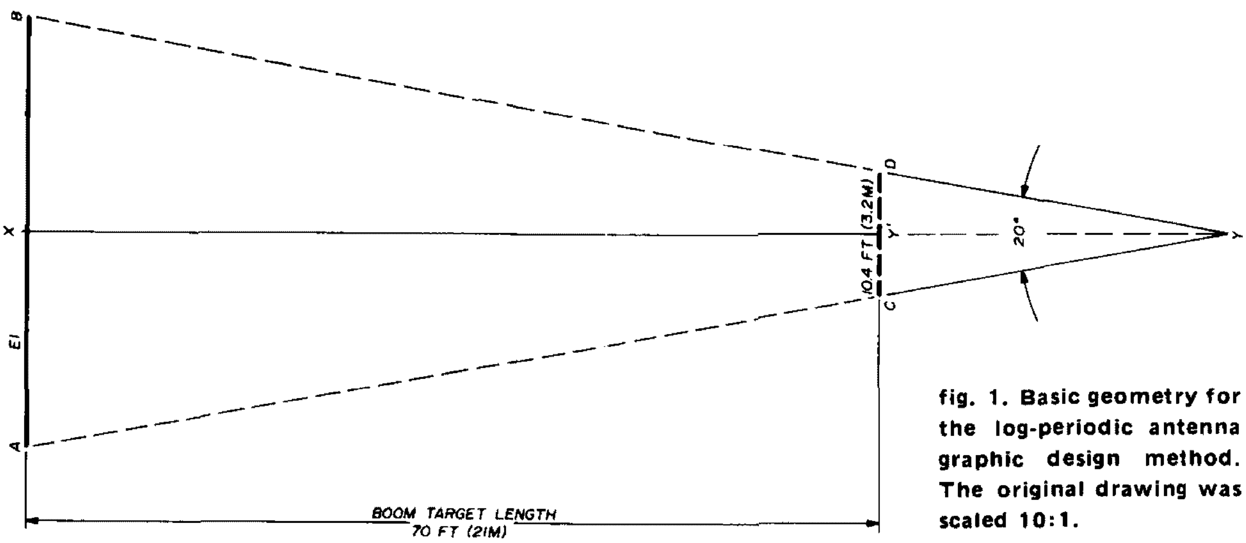
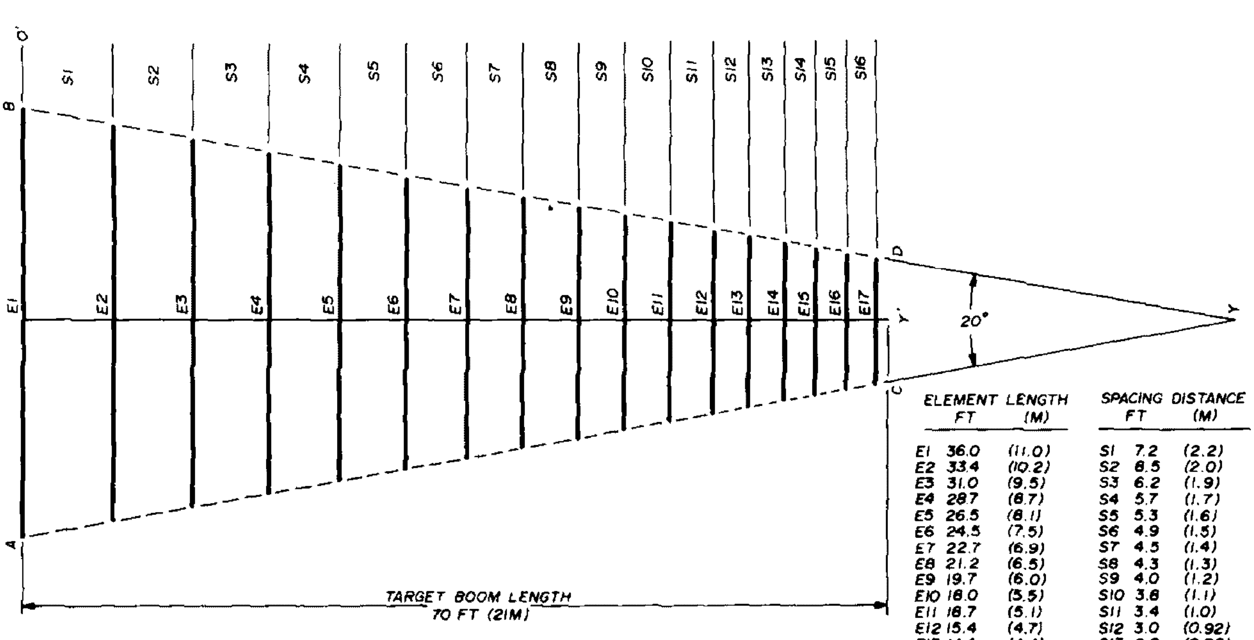


fig. 1. Basic geometry for the log-periodic antenna graphic design method. The original drawing was scaled 10:1.

In addition, several special log-periodics have been designed on paper covering other frequencies. Some covered both MARS and amateur bands; several were for special vhf and uhf TV channels; and, I blush to say, one covered 26-27 MHz for a CBer wanting a good skip antenna.

After designing my first three log-periodics the hard way with the formulas, I felt there must be an easier design method. When designing the



ELEMENT LENGTH	FT	(M)	SPACING DISTANCE	FT	(M)
E1	36.0	(11.0)	S1	7.2	(2.2)
E2	33.4	(10.2)	S2	6.5	(2.0)
E3	31.0	(9.5)	S3	6.2	(1.9)
E4	28.7	(8.7)	S4	5.7	(1.7)
E5	26.5	(8.1)	S5	5.3	(1.6)
E6	24.5	(7.5)	S6	4.9	(1.5)
E7	22.7	(6.9)	S7	4.5	(1.4)
E8	21.2	(6.5)	S8	4.3	(1.3)
E9	19.7	(6.0)	S9	4.0	(1.2)
E10	18.0	(5.5)	S10	3.6	(1.1)
E11	16.7	(5.1)	S11	3.4	(1.0)
E12	15.4	(4.7)	S12	3.0	(0.92)
E13	14.4	(4.4)	S13	2.9	(0.88)
E14	13.3	(4.1)	S14	2.7	(0.82)
E15	12.4	(3.8)	S15	2.5	(0.76)
E16	11.6	(3.5)	S16	2.3	(0.70)
E17	10.4	(3.2)			

fig. 2. Log-periodic antenna using 17 elements and a 69-foot (21m) boom. Approximate gain (referenced to a dipole) for this design and those in figs. 3 to 5 is 10 dB. All cover 14 to 30 MHz.

original antennas, I always made an assembly sketch on graph paper after arriving at the correct element lengths and spacing distances. Since the outline of a log-periodic results in an isosceles triangle, with the long rear element being the base and the shorter forward elements forming the triangle toward the apex, the following simple no-math graphic design method became apparent. I believe this simple design method will be of interest to any amateur wishing to design a log-periodic for a particular band, bandwidth, or to fit a log-periodic into a given space.

graphic design method

You will need the following materials: graph paper, 1/10 cross section, 8½x10½ inch (21.5x26.7 cm) or larger; an architect or engineer's scale; a protractor; and some French curves (not absolutely necessary but helpful in designing the side catenary lines).

For the first example we will design an L-P for 20, 15 and 10 meters or for operation on any frequency between 14 and 30 MHz.

1. First determine the low- and high-frequency cutoff required or frequencies over which the L-P is to operate, or its bandwidth.

2. Next determine the amount of space available when the L-P is aimed in the desired direction. If there is a space limitation, it may be necessary to reduce the boom length, losing some gain. This is discussed later.

3. Determine the length of the longest (rear) element and the shortest (forward) element:

Rear element. The rear element should be at least 5% longer than the lowest cutoff or operating frequency. Using the usual formulas:

$$\text{½ wavelength} = 468/\text{MHz} = 468/14 = 33.4 \text{ feet } (10.2\text{m})$$

$$33.4 + 5\% \approx 34.4 + 1.7 \approx 35.1 \text{ feet } (10.7\text{m})$$

Since a slightly longer length is better, we will use 36 feet (10.9m) for the rear

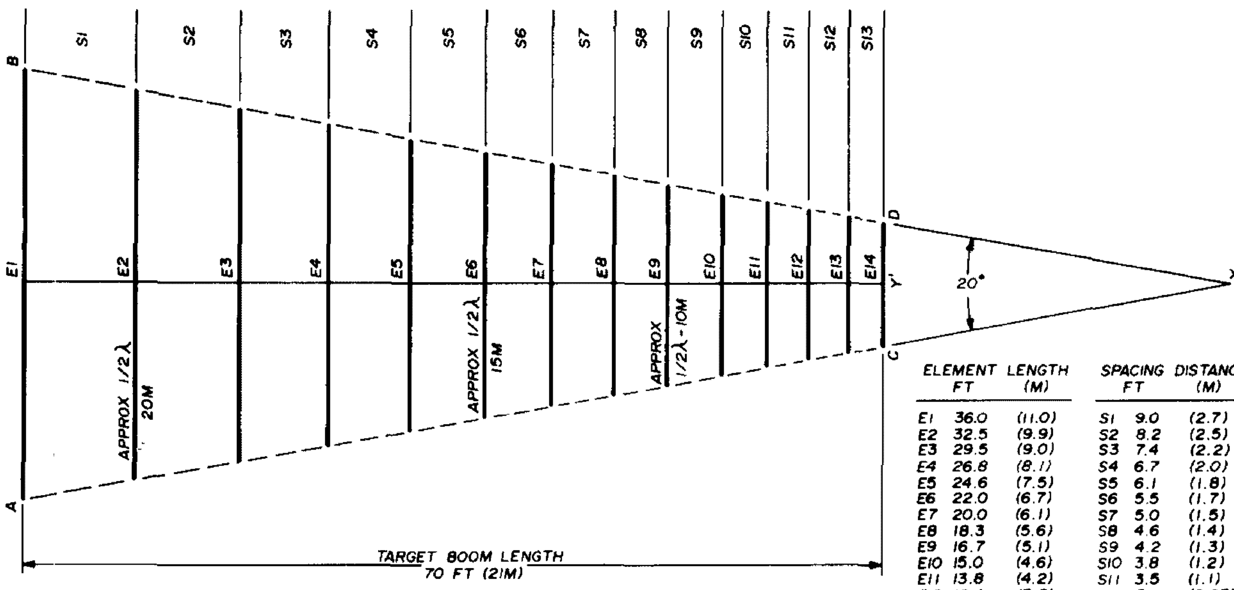


fig. 3. An L-P with 14 elements and a 70-foot (21.3m) boom.

ELEMENT	LENGTH FT	(M)
E1	36.0	(11.0)
E2	32.5	(9.9)
E3	29.5	(9.0)
E4	26.8	(8.1)
E5	24.6	(7.5)
E6	22.0	(6.7)
E7	20.0	(6.1)
E8	18.3	(5.6)
E9	16.7	(5.1)
E10	15.0	(4.6)
E11	13.8	(4.2)
E12	12.4	(3.8)
E13	11.4	(3.5)
E14	10.4	(3.2)

SPACING	DISTANCE FT	(M)
S1	9.0	(2.7)
S2	8.2	(2.5)
S3	7.4	(2.2)
S4	6.7	(2.0)
S5	6.1	(1.8)
S6	5.5	(1.7)
S7	5.0	(1.5)
S8	4.6	(1.4)
S9	4.2	(1.3)
S10	3.8	(1.2)
S11	3.5	(1.1)
S12	3.1	(0.93)
S13	2.9	(0.87)

element. This element length will resonate at 468/36 = 13.0 MHz.

Forward element. The shortest element should resonate 45 to 50% higher in frequency than the desired high-frequency cutoff. From my experience, the swr will be lower by using a high-frequency cutoff plus at least 50%:

30 MHz + 50% = 45 MHz

½ wavelength at 45 MHz = 468/45
= 10.4 feet (3.2m)

We now have the required length of the rear element, 36 feet (10.9m) and the forward element, 10.4 feet (3.2m)

4. We will now estimate a boom target length to determine a practical distance from the long rear element (E1) to the short forward element. From experience, an L-P designed to cover an octave (2:1 bandwidth) should have a boom length from 1.5 to 3 times the length of the rear element. If the boom length is less than 1.5 times the rear-element length, the apex angle will exceed 40 degrees and the forward gain will suffer. In other words, the gain drops off quite rapidly for a boom length less than E1 x 1.5, or with an apex angle of more than 40 degrees.

From the L-P formulas and nomographs in the references it will be noted that the α angle (which is 1/2 the apex angle); relative spacing, σ; design or scale factor, τ; and other variables all govern the forward gain, front-to-back ratio, etc. The design factor, τ, given by references 4 and 5, is of special interest because it gives gain figures between 7.5 to 12 dB for various combinations of these formulas. However, the purpose of this article is to eliminate all formulas, so the information above is for those wishing to pursue further study.

We now have sufficient dimensions to start drawing the graphic L-P. In laying out the first antenna, use a scale of 10:1. Referring to fig. 1, proceed as follows:

5. First draw the longitudinal center line, X-Y. Now draw the longest element, E1, (line A-B) determined by **step 3** to be 36 feet (10.9m).

6. A boom length of 2 x E1 will be used for the first example; 36 x 2 = 72 feet (21.6m). We will use 70 feet (21m)

as the target length and try not to exceed this length.

It will be found later that this overall length may vary plus or minus a few feet, but we will try not to exceed an overall length of 70 feet (21m). This will also give us a target length for fore and aft mast spacing.

Now measure 70 feet (21m) along the X-Y centerline from point X (or the rear element) toward Y, making a dot at exactly 70 feet (21m), which will be called Y′.

7. Next draw a temporary (dotted) line, C-D, at the 70 foot (21m) point, making this line 10.4 feet (3.1m), which is the length of the shortest (forward) element. Make certain that the X-Y axis bisects this line, or that the two ends of line C-D are equidistant from X-Y. A permanent line is not drawn here as this is a temporary, or target line. It is a step of the graphic method needed to generate the *triangle outline* to which the other elements will be added. The final, short element, may not coincide exactly with the boom target length. It may miss this distance by a few feet; however, this has little effect on the performance of the antenna.

8. Draw line A-C and extend it until it crosses the X-Y axis. Next draw line B-D, extending it to also cross X-Y. If all drawings to this point have been accurate, these lines will meet, forming the apex, (Y), of an isosceles triangle, A-Y-B.

9. The next step is to add the remaining elements between the rear element, E1, (line A-B) and the short, forward, target element, which should fall near the temporary element (line C-D).

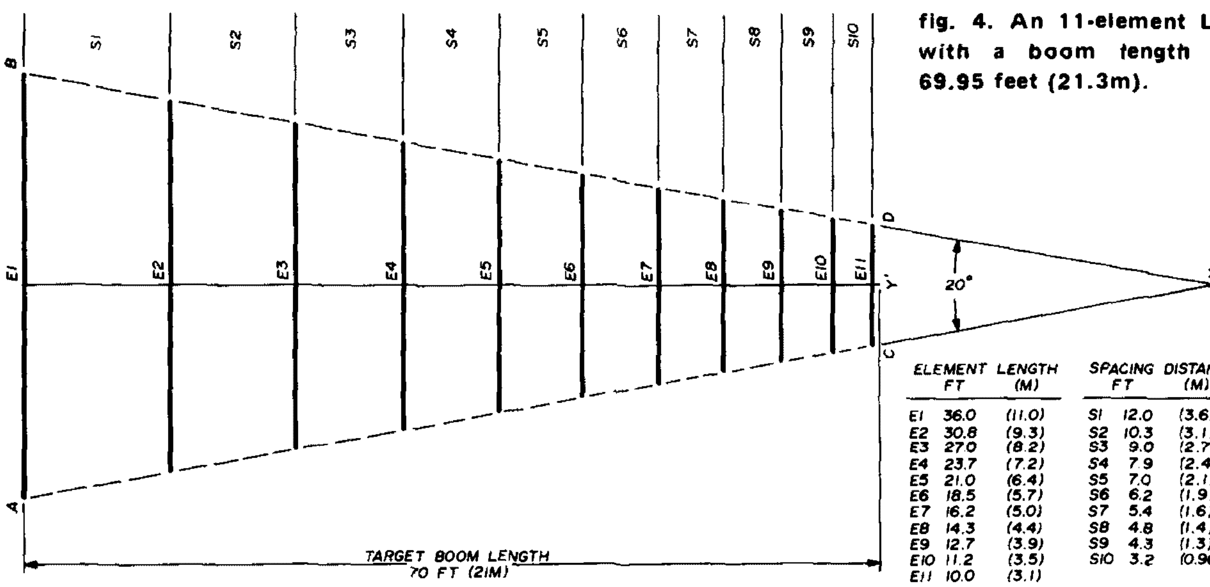


fig. 4. An 11-element L-P with a boom length of 69.95 feet (21.3m).

ELEMENT	LENGTH FT	(M)
E1	36.0	(11.0)
E2	30.8	(9.3)
E3	27.0	(8.2)
E4	23.7	(7.2)
E5	21.0	(6.4)
E6	18.5	(5.7)
E7	16.2	(5.0)
E8	14.3	(4.4)
E9	12.7	(3.9)
E10	11.2	(3.5)
E11	10.0	(3.1)

SPACING	DISTANCE FT	(M)
S1	12.0	(3.6)
S2	10.3	(3.1)
S3	9.0	(2.7)
S4	7.9	(2.4)
S5	7.0	(2.1)
S6	6.2	(1.9)
S7	5.4	(1.6)
S8	4.8	(1.4)
S9	4.3	(1.3)
S10	3.2	(0.96)

Precise measurements must be made for the remainder of this design. For this reason an accurate scale must be used and the lengths should be read or estimated to 0.1 foot (3cm). **Fig. 2** will now be used to complete the L-P. Our next objective will be to determine a correct spacing ratio between the elements.

element spacing

Since a log-periodic antenna can be considered a unidirectional, end-fire array having a series of driven elements, a spacing distance of 0.05 to 0.2 wavelength should provide the best gain, as

the two adjacent elements are out of phase because of transposition between elements, which is required for a log-periodic antenna.

For starters we'll use an element spacing of approximately one-tenth wavelength between elements such as the rear (longest) element, E1, and the following element, E2; and between E2 and E3, etc. An easy method of approaching this spacing is to divide the triangle. As shown by fig. 2 this length will be 33.4 feet (10.2m). (The remaining elements and element spacing will be referred to as E3...En and S2...Sn, respectively.)

11. Next determine the second spacing distance, S2: E2/5 = 32.5/5 = 6.5 feet (2.0m). Mark this distance, S2, draw E3, and measure its length, which will be 31.0 feet (9.4m).

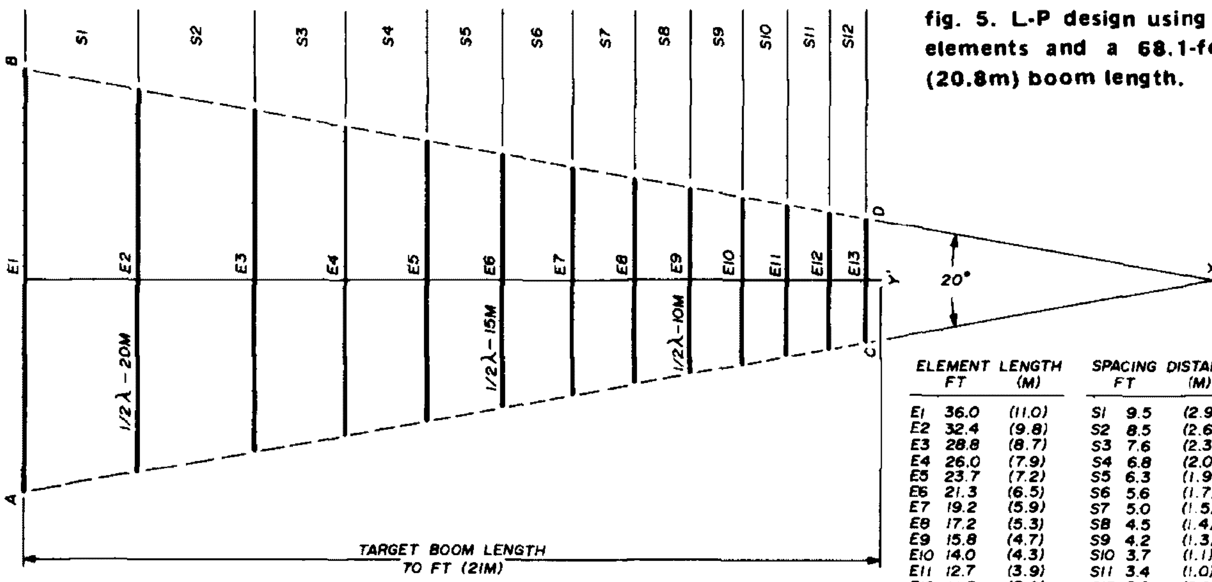


fig. 5. L-P design using 13 elements and a 68.1-foot (20.8m) boom length.

ELEMENT	LENGTH FT	(M)	SPACING	DISTANCE FT	(M)
E1	36.0	(11.0)	S1	9.5	(2.9)
E2	32.4	(9.8)	S2	8.5	(2.6)
E3	28.8	(8.7)	S3	7.6	(2.3)
E4	26.0	(7.9)	S4	6.8	(2.0)
E5	23.7	(7.2)	S5	6.3	(1.9)
E6	21.3	(6.5)	S6	5.6	(1.7)
E7	19.2	(5.9)	S7	5.0	(1.5)
E8	17.2	(5.3)	S8	4.5	(1.4)
E9	15.8	(4.7)	S9	4.2	(1.3)
E10	14.0	(4.3)	S10	3.7	(1.1)
E11	12.7	(3.9)	S11	3.4	(1.0)
E12	11.5	(3.6)	S12	3.0	(0.9)
E13	10.2	(3.2)			

rear-element length by 5: 36/5 = 7.2 feet (2.2m). For this simple design method, this might be considered similar to the "relative spacing ratio, σ, = dN/21n," or "the mean spacing factor, σ', = dn/21n;" or "design ratio, τ, = 1n/L1-1;" in the log-periodic design formulas, which can be obtained from references 1-5 and 11. We shall, however, proceed with our no-math design method.

10. Refer to fig. 2 and accurately measure 7.2 feet (2.2m) from the center of element E1 (point X) down X-Y and mark the position with a dot. This point will be the location of the second element, E2. Now draw E2 and measure its exact length between the sides of the

12. Continue the mark-draw-measure-divide procedure, very accurately, to obtain the remaining spacing distances and element lengths until the last and shortest element is reached, which should be close to 10.4-feet (3.2-m) long and should be approximately ±1 foot (30cm) from our boom target length, depending on how accurately the measurements have been made and drawn. We now have all element lengths and spacing distances for a 17-element log-periodic for 14-30 MHz having a boom length of approximately 69 feet (21m). As measured by the protractor, the apex angle is approximately 20 degrees. Considering this angle, number of elements, and the boom length, this

L-P should give a forward gain of approximately 10 dB if the antenna is one-half wavelength above ground, or at least 35 feet (10.7m) on 20 meters.

Although 17 elements are good from an swr standpoint, a smaller number of elements can be used, as mentioned later. Twelve to 13 elements are generally ratio L(E)/3 is used, resulting in only 11 elements. As 12 to 13 elements are a minimum for a 2:1 bandwidth, **fig. 5** is generated using a ratio L(E)/3.8, which gives 13 elements. This should be the optimum of the four L-Ps all having approximately the same final boom length and apex angle.

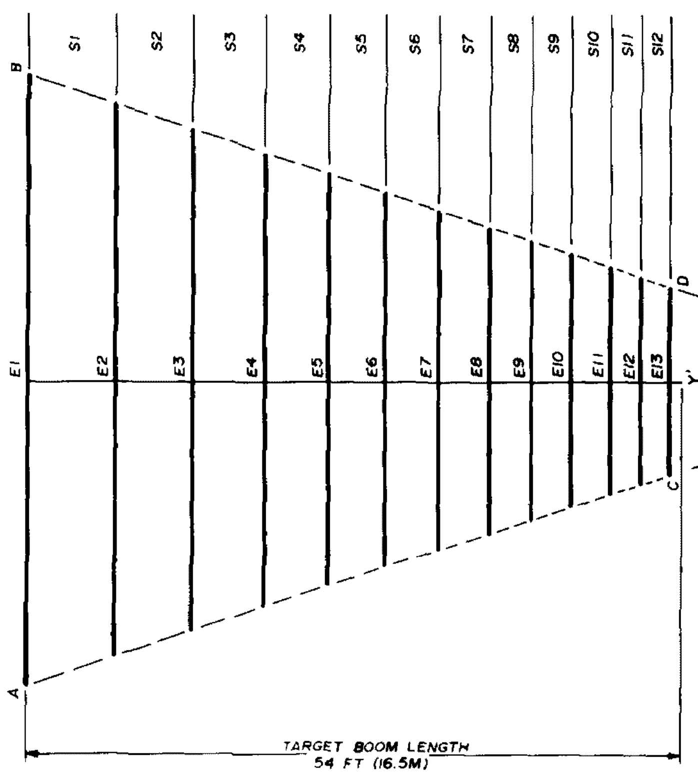


fig. 6. Same design as in fig. 5 except with minimum boom length (53 feet or 16.2m). Gain is approximately 8 dB.

ELEMENT	LENGTH FT	(M)	SPACING	DISTANCE FT	(M)
E1	36.0	(11.0)	S1	7.2	(2.2)
E2	32.5	(9.9)	S2	6.5	(2.0)
E3	29.4	(8.9)	S3	5.9	(1.8)
E4	26.8	(8.1)	S4	5.4	(1.6)
E5	24.3	(7.4)	S5	4.8	(1.4)
E6	22.0	(6.7)	S6	4.4	(1.3)
E7	20.0	(6.1)	S7	4.0	(1.2)
E8	18.0	(5.5)	S8	3.6	(1.1)
E9	16.3	(5.0)	S9	3.2	(0.96)
E10	14.8	(4.5)	S10	3.0	(0.90)
E11	13.2	(4.1)	S11	2.6	(0.78)
E12	12.0	(3.7)	S12	2.4	(0.72)
E13	10.6	(3.3)			

the minimum required for an L-P designed to cover an octave. Using less than 17 elements will reduce weight, cost and labor.

a 14-element design

Fig. 3 illustrates a similar 14-30 MHz L-P also using a 70-foot (21.3m) boom target length but instead of the ratio L(E)/5, a ratio of L(E)/4 is used, resulting in 14 elements. Note that the boom length is almost exactly 70 feet (21.3m) and the apex angle remains the same as the 17-element L-P, **fig. 2**.

11- and 13-element designs

Fig. 4 is an equivalent L-P but the

Any one of these designs would have about equal forward gain; however, the 11-element design would probably not have as smooth or flat an swr across its bandwidth due to a minimum number of elements vs bandwidth. The swr might exceed 2:1 on some frequencies.

The 13-element array is one of the L-Ps assembled and being used here. Note that an odd number of elements is suggested from a mechanical assembly standpoint, as explained in some of my previous articles.

From the four L-Ps illustrated in **fig. 2, 3, 4** and 5, it will be noted that the elements vary as follows: 11 for L(E)/3, 13 for L(E)/3.8, 14 for L(E)/4, and 17

for L(E)/5, each having essentially the same boom length and apex angle. Each should give approximately 10 dB gain.

14-30 MHz log-periodic with minimum boom length

If space is not available for a 70-foot (21.3m) boom length, the minimum length of 54 feet (16.5m) (E1 x 1.5) can be assembled per **fig.** 6. The ratio of L(E)/5 is used, which gives 13 elements. Since this antenna has a length only 1.5 times that of the rear element (or approximately 3/4 wavelength boom length), and the apex angle is 36 degrees, its gain will probably not exceed 8 dB.

14-30 MHz L-P with 100 foot (30.5m) boom length

For those desiring maximum gain from an L-P for 20, 15 and 10 meters and if space is available, the 14-30 MHz L-P boom length can be extended to approximately 100 feet (30.5m) as illustrated by **fig.** 7 which, if properly assembled and suspended at least 40 feet (12.2m) above ground, will give a gain of 12 dB.

L(E)/3.3 was used for generating this L-P. The boom target length was 100 feet (30.5m). For this drawing the last element, E17, is 101 feet (30.82m) from the rear element starting point, which overshot our target by 1.1 foot (34cm). Sixteen elements could be used, which would be 98.0 feet (29.9m), but as mentioned previously, an odd number of elements is desirable from a mechanical standpoint; therefore, the extra 1.1 feet (34cm) should be acceptable.

gain vs boom length and apex angle

By the graphic design method we have generated three 14-30 MHz dipole log-periodic antennas having three different boom lengths, apex angles and gain:

boom length feet	boom length meters	apex angle (degrees)	approximate gain (dB)
54	16.5	36	8
70	21.3	20	10
100	30.5	15	12

Thirteen elements are suggested for the shorter 50- and 70-foot (15.2 and 21.3m) L-Ps and 17 elements for the 100-foot (30.5m) length. I have tried all three configurations. A 70-foot (21.3m) L-P is used for my northeast beam and the 100-foot (30.5m), 17-element array for the beam directed west, which has given outstanding performance.

Since the 10-meter band may not be of interest now due to propagation conditions, this portion of the L-P designs can be eliminated by deleting elements shorter than 15 feet (4.6m); i.e., for the 13 element L-P, **fig.** 5, elements 10, 11 12 and 13 can be deleted, reducing the length by 14.3 feet (4.4m) leaving a 9-element L-P covering 14-21.5 MHz for 20 and 15 meters.

For the shortest L-P, **fig.** 6, deleting the four forward elements would reduce the length by 10.9 feet (3.3m) or a boom length of 41.2 feet (12.6m). This 20- and 15- meter beam could then be erected in a space 40 x 45 feet (12.2x13.7m). Likewise, for the 100-foot (30.5-m) L-P, **fig.** 7, elements 12, 13, 14, 15, 16 and 17 can be deleted, reducing its length 23 feet (7m), leaving an 11-element L-P with boom length of 78.2 feet (23.9m). The deletion of the 10-meter section of these L-Ps will have little effect on the gain on 20 and 15 meters, since the apex angle is unchanged.

40- and 20-meter log-periodics

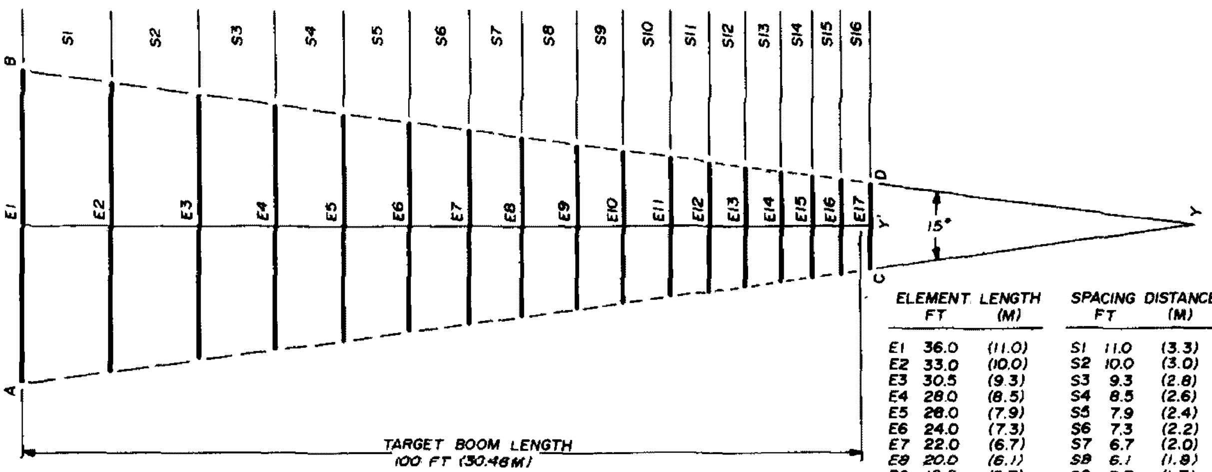
The same graphic approach can be used for designing an L-P covering 7-14.5 MHz for operation on 40 and 20 meters; however, by doubling the element lengths, spacing distances, and boom length of the 20-, 15- and 10-

meter L-Ps (fig. 2 through 7), 40- and 20-meter L-Ps can be assembled, except possibly that of fig. 7, since it would be over 200 feet (61m) long, The L-Ps of fig. 2, 3, 4 or 5 would become 140 feet (42.7m) long; however, the shortest L-P, (fig. 6), would be only 108 feet (32.9m) long.

By adding four or five short forward elements and extending the boom length slightly, a 40- and 20-meter L-P can be modified to cover 7 to 21.5 MHz for operation on 40, 20 and 15 meters, which are probably the most-used bands at present. A 40-, 20- and 15-meter L-P was described in reference 7, which was of the skip-band type having a portion deleted between 7 and 14 MHz. An L-P for 40 meters should be at least 70 feet (21.3m) high for DX work.

For those wishing to design an L-P covering a bandwidth greater than a single octave, 12 to 13 elements are about the minimum that should be used. Additional elements can be used as the boom length is increased. For a 3:1 bandwidth, no less than 18 or 19 elements will be required; for 4:1 approximately 21 will be required.

In reference 7 a single-band L-P for 40 meters was described which has been tested here. Single-band L-Ps can be assembled for any of the high-frequency bands using the graphic design method. Five elements are sufficient and the swr remains relatively flat across the band for which the antenna is designed. A total of five different single-band L-Ps have been tested here on several bands. With an apex angle of 32 - 36 degrees, the single-band L-P will generally show a gain of 8 to 10 dB provided it is suspended approximately one-half wavelength above ground.



ELEMENT	LENGTH FT	(M)	SPACING	DISTANCE FT	(M)
E1	36.0	(11.0)	S1	11.0	(3.3)
E2	33.0	(10.0)	S2	10.0	(3.0)
E3	30.5	(9.3)	S3	9.3	(2.8)
E4	28.0	(8.5)	S4	8.5	(2.6)
E5	26.0	(7.9)	S5	7.9	(2.4)
E6	24.0	(7.3)	S6	7.3	(2.2)
E7	22.0	(6.7)	S7	6.7	(2.0)
E8	20.0	(6.1)	S8	6.1	(1.8)
E9	18.5	(5.7)	S9	5.7	(1.7)
E10	12.0	(3.7)	S10	5.2	(1.6)
E11	15.8	(4.8)	S11	4.8	(1.5)
E12	14.5	(4.5)	S12	4.4	(1.3)
E13	13.3	(4.1)	S13	4.0	(1.2)
E14	12.2	(3.8)	S14	3.7	(1.1)
E15	11.3	(3.5)	S15	3.4	(1.03)
E16	10.4	(3.2)	S16	3.1	(0.94)
E17	9.5	(3.0)			

fig. 7. Big Daddy — an L-P with 17 elements and a boom length of 101.1 feet (30.8m). Gain is approximately 12 dB.

vertical monopole log-periodic antennas

The vertical monopole L-Ps for 40 and 80 described in reference 9 were also assembled by the graphic design method. The ratio L(E)/5 is best for a single-band, 5-element dipole or monopole L-P since element E2 will be spaced approximately one-quarter wavelength from the shortest element; and the open-wire center feeder, which is one-quarter wavelength from the feedpoint to E2, or the active element, also serves

as an impedance-matching transformer between element E2 and the higher impedance at the feedpoint, which is of the order of 200 - 300 ohms. Thus a 4:1 balun can be used.[12]

summary

Of the nearly twenty different log-periodic antennas I have built and tested, all but the first three (erected in 1970) have been designed by the graphic method. Although this simple procedure may seem crude, and may not be as accurate as an L-P designed entirely by the formulas, all antennas have produced the same results. In addition to the dipole log periodics, I have erected and tested three of the monopole L-Ps, which have quarter-wavelength vertical radiators and ground radials. These were tested on 40 and 80 meters. My graphic design method has also been applied to two of the trapezoidal sawtooth and zig-zag log-periodic designs now being tested on 20 and 15 meters.

I hope the simple non-math graphic design method will be of help to those wanting a special L-P for a particular frequency range or to fit in a limited space. To obtain maximum gain, make the L-P as long as possible for the given space; this will in turn give a minimum apex angle. Also use a sufficient number of elements to keep the swr relatively flat across the band for which the antenna is designed.

If you plan an L-P for 20, 15 and 10 meters, try to make the boom at least 54 feet (16.5m) long and use a minimum of 13 elements. If you are not interested in the 10-meter band, it would be better to design the L-P to cover 14 to 22 MHz only for 20 and 15 meters, using the 54-foot (16.5-m) boom length for a 9-element 14 to 22 MHz L-P. This would reduce the apex angle to 32 degrees and the gain would be approximately 10 dB vs 8 dB for the 20-, 15- and 10-meter antenna on the same boom length.

acknowledgements

I wish to thank the many amateurs who have assisted in on-the-air tests of these L-Ps for the past four years and for the many letters, phone calls, visits and QSOs by those interested in L-P design. I particularly thank those who have actually erected and are now using L-Ps similar to mine described in the references.

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ham radio

four-element phased vertical array

Design and construction of a four-element phased array for 40 meters that provides nearly 9 dB forward gain

Jerry Swank, W8HXR, 657 Willabar Drive, Washington, Ohio 43160

For the last ten years, during the Antarctic winter, I have been handling phone-patch traffic from KC4-land on 40 meters. At this end I have been using a pair of phased verticals. In Antarctica they have used log periodics, vees, rhombics and monopoles. I usually receive good signal reports, but their signals are often buried in summer static and difficult to copy.

I have often tried to get the operators in Antarctica to try phased vertical antennas, but they change crews every year, and no one was apparently very interested. Then, three years ago, I began working Byrd Station nearly every night with Bob Conner, and I told him that I thought a set of phased vertical antennas would solve their problem.

When he came back for two-years stateside duty we got together and planned that when he went back to Antarctica in 1974 we would try out the idea. I outlined the plan for the vertical array, and we discussed it thoroughly, planning to ship everything to Antarctica, down to the last solder lug.

The Navy bought four trapped vertical antennas, Hy-Gain 14AVQs, and Bob got the necessary wire and fasteners to mount them. The logistics turned out to be a real cliff hanger with the antennas barely making it on board the last plane into McMurdo Station. It was a great deal of work in the icy weather, but finally, on February 9th, 1974, one vertical was installed. Bob was pretty discouraged, though, because the single vertical picked up so much noise from the big, nearby Navy installation that it was unusable. I told him to be patient and put up another one a half-wavelength away and hook them up as a broadside array. On February 11th the signals from McMurdo came through loud and clear, and Bob said that the interference had dropped almost to zero.

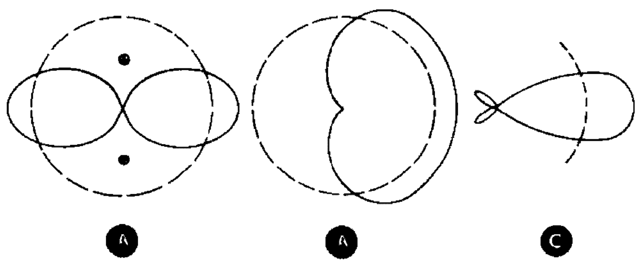


fig. 1. Horizontal radiation pattern of half-wavelength spaced vertical antennas is shown in (A). Pattern at (B) is that of quarter-wavelength spaced verticals. At (C) is pattern of the four-element phased array discussed here.

By arranging the location of a phased array and properly orienting it, noise can be cut as much as 40 dB. In fact, this was the original reason I had gone to phased verticals. The noise from the 7200-volt utility line at my back property line, 12 feet (3.7 meters) from the antennas, put a steady buzz into my receiver at about S6. When I installed the phased array, the noise dropped practically to zero.

When the four vertical antennas were installed at McMurdo Station, their 40-meter signals began coming in about 0700 GMT at 10 to 40 dB over S9. Everyone there (and here) was ecstatic, and stations all over the states began breaking-in to tell the KC4USV operators their signals had never been so loud. Note that this was before the 40-meter season normally opened. We did not expect to use 40 meters except for tests, and had planned to meet on 20 meters for coordination. We did not expect 40 to be usable until sometime in April.

the array

The broadside array with half-wavelength spacing has very broad nulls on the side and a narrow 60° beam which takes out noise on either side quite well (fig. 1A). The forward gain is about 3.86 dB. Quarter-wavelength spacing of the verticals results in a cardioid pattern (fig. 1B); the beam width is about 120° and the only good null is toward the rear. However, this pattern has a lower forward angle of radiation, and is best for extreme distances. Forward gain is about 4.5 dB.

Combining the half-wavelength spaced broadside array with the quarter-wavelength spaced system (fig. 2) provides nearly 9 dB forward gain. This is better than a three-element beam and at a lower angle unless you have a real man-sized tower. The pattern is similar to that of a Yagi (fig. 1C).

construction

A plan view of the four-element phased array is shown in fig. 3. The phasing lines between the quarter-wavelength spaced elements (A-D and B-C) are 3/4-wavelength long because,

when the velocity factor of the RG-8/U coaxial cable (0.66) is taken into consideration, a quarter-wavelength coaxial line is too short to reach between the elements. The 3/4-wavelength phasing line provides the same phase delay.

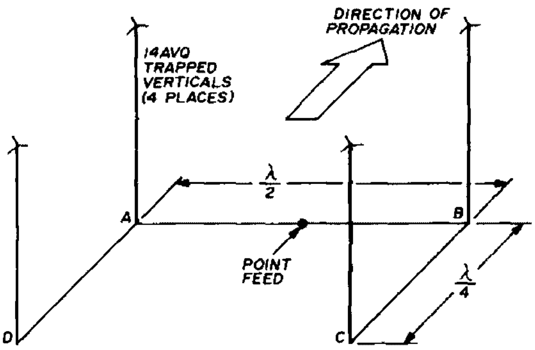


fig. 2. The four-element 40-meter phased array used at KC4USV uses four Hy-Gain 14AVQ trapped verticals.

Although a tee connector is shown between the two half-wavelength spaced elements (A and B), in the installation at McMurdo Station the tee connector is actually located inside the station. Because of the severe weather conditions during the winter, and frequent 100 mph (160 kmh) winds, this system allows either quarter-wave spaced pair to be used separately in case one of the verticals of the other pair is brought down or damaged by the wind.

It would also be possible to feed antennas C and D with one feedline, and antennas A and B with another, placing the phasing line inside the shack. With this system the pattern of the array could be reversed.

The antennas at McMurdo Station are mounted on a hill of volcanic ash and the electrical ground is down about 30 feet (9m). It was originally planned to use ground radial mounting, but the volcanic ash has extremely low conductivity, and the antennas are mounted on about six feet (1.8m) of pipe, so they are located a quarter wavelength above electrical ground. This is ideal for a ground plane. One radial was installed on the far side and five radials on the near side to support the antennas and provide the best ground plane toward the desired direction (see fig. 3). The radials are each insulated at the outer, high-voltage ends.

performance

Because of poor performance in previous years, in 1974 the Navy installed a Collins log-periodic on a 100-foot (30.5m) tower. We tested the phased array twice against the big log-periodic, once under good conditions, and once under the poorest conditions. The log-periodic had less than 3 dB advantage over the phased vertical array. The old 40-meter beam, used for years, sounds pitiful by comparison.

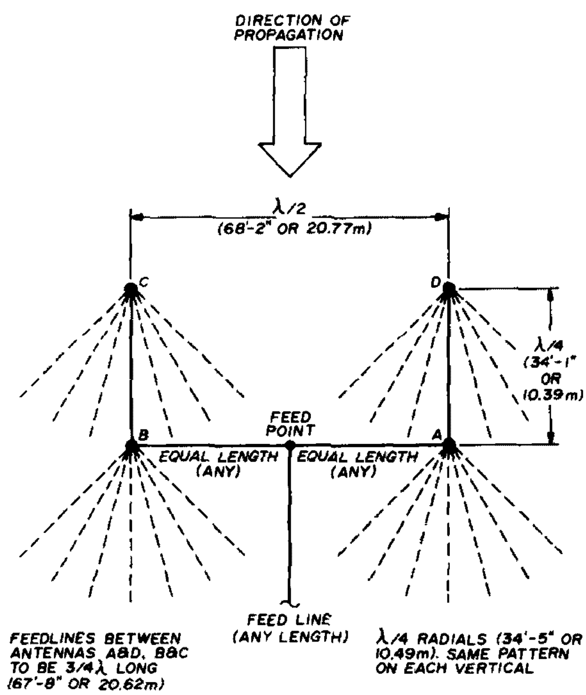


fig. 3. Dimensions of the 40-meter phased array. Gain of this antenna is nearly 9 dB.

An unexpected bonus made the phased array even more valuable. Since the four verticals are trapped designs for operation on 40, 20, 15 and 10 meters, performance on 20 meters is good. By

dropping one cardioid section, the remaining section becomes a half-wavelength end-fire array on 20 (it also works well on 15 meters). Trapped antennas were originally chosen because they were shorter, and less likely to be blown down in high winds, so multiband operation of the array was an added bonus.

The phased array at KC4USV is mounted approximately 6 feet (1.8m) off the ground on a sloping hillside so the rear pair of antennas is about 7 feet (2.1m) higher than the forward pair. This results in an extremely low vertical radiation angle. Standing by one of the antennas is Bob Conner, WB2BCJ, who installed the array at McMurdo Station in Antarctica.

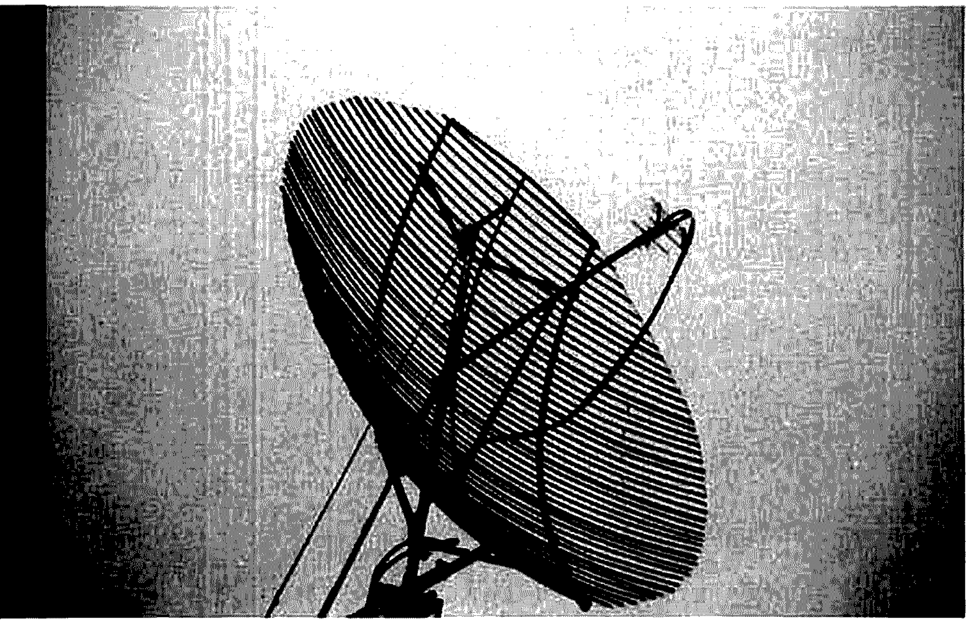
postscript

On July 22nd, 1974, an ice storm hit McMurdo Sound with winds to 138 mph (222 kmh), tearing off the roof and one wall of the ham shack. The storm also took out the large elements of the big log periodic, making it unusable on 40 meters. The four 14AVQ verticals in the array, however, were completely undamaged.

All the amateur radio gear was moved to another building and two spare 14AVQ vertical antennas were installed temporarily as a half-wavelength spaced broadside pair so contact was resumed in a few days. When the ham shack was rebuilt the four-element phased array was placed back in operation. The big log-periodic on the 100-foot tower, on the other hand, had to wait until summer.

As a sidelight, to show what the weather is like at McMurdo Station, on August 8th a plane came in from Christ Church, New Zealand, 2000 miles away, air dropped nine sacks of mail, weighted with sand, from about 400 feet (122m) and returned non-stop to New Zealand. One bag broke open and three others were carried away by the wind. Two of the three bags were found 20 miles (32km) away and one was finally recovered 45 miles (72km) away. The mail from the broken bag was only partially recovered!

ham radio



parabolic reflector

element spacing

It's not necessary to use solid or screened reflector surfaces — properly spaced reflector elements provide the same performance

Norman J. Foot, WA9HUV, Elmhurst, Illinois 60126

The subject of parabolic reflector element spacing is like the weather — nearly everybody talks about it, but few people do anything about it. One reason is that there are a large number of parameters relating to good parabolic antenna design, and the matter of reflector element spacing seems to have been lost in the shuffle.*

It is possible to avoid the issue of reflector spacing by adding metal screening or using a solid metal surface as many builders have done.[1,2] Some of these designs however, take special skills and the cost is likely to be high. As an alternative, an all-metal dish can be purchased on the surplus market.

*Sometime after preparing the material for this article, author WA9HUV discussed it with Edward F. Harris, Chief Engineer with Mark Antennas, and discovered that Harris had done similar work in the 1950s. His report, "Designing Open-Grid Parabolic Antennas," appeared in the November, 1956, issue of *Electronic Industries* magazine.

On the other hand, several low cost parabolic-reflector designs employing aluminum tubing have appeared in amateur radio magazines in recent years.[3,4] The question is, are these designs really inferior to solid reflector designs, and if so, to what extent? What are the performance compromises one is willing to make when consideration is given to cost, mechanical strength and stability, portability, ease of construction, mounting, wind-loading and weight?

It seems reasonable to assume that if the elements of a reflecting screen are located extremely close together in terms of operating wavelength, then most of the incident energy will be reflected and very little will be transmitted through the screen. On the other hand, if the reflector elements are many wavelengths apart, it seems clear that a large portion of the incident energy should pass through uninhibited. By intuition, it does not seem likely that a screen of reflectors suddenly becomes completely reflective at some critical value of element spacing. It also follows that the screen probably does not become completely transparent when this critical value of element spacing is slightly exceeded. Just how does the reflective property vary with element spacing?

In an attempt to answer these questions I performed a series of experiments. The results, which are the subject matter for the balance of this article, should be of interest to future builders of parabolic reflectors.

test setup

To perform the experiments, a wooden frame 20-inches (51-cm) wide was made of furring strips. Holes were drilled in the sides of the frame 3/4-inch (19-mm) apart so that 24-inch (61-cm) long aluminum tubes, 7/16-inch (11-mm) OD, could be inserted and stacked like a venetian blind, with the elements separated by 3/4 inch or any multiple of 3/4 inch. A sketch of the frame is presented in fig. 1.

Two kinds of experiments were performed, one to test the transmission and the other the reflective properties of the screen as element spacing was varied. A

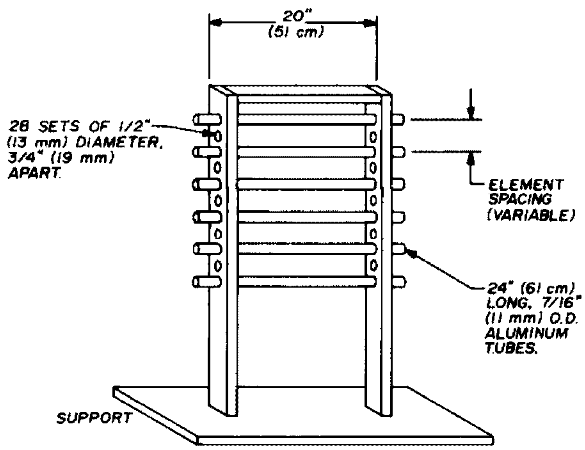


fig. 1. Sketch of wooden frame used to vary the space between reflector elements. Test setups are shown in fig. 2 and 3.

battery-operated 2304-MHz small-signal source driving a circular horn antenna was used as the transmitter.

A similar horn antenna was used to feed a 2304-MHz converter. The 2304-MHz test frequency was chosen so that the reflecting screen could be kept reasonably small. If the experiment were performed at 432 MHz, for example, much longer reflector elements and a physically larger frame would have been required.

In the second set of experiments both the receiving and transmitting horn antennas were located on the same side of the reflector, side-by-side, and oriented so that maximum reflected signal was received. When the reflector was removed, the crosstalk between horns was too small to operate the S-meter in my receiver. The elements of the screen were then added in various combina-

tions of spacing, and the S-meter reading recorded for each case.

test results

In the first set of experiments in which the receiving and transmitting antennas were separated by the screen, very little signal reached the receiving antenna for element spacings up to 1½ inches (38mm). At 2¼ inch (57mm) spacing, the screen became slightly opaque, while at 4 inches (102mm) it was almost completely transparent. It is interesting that for the 3-inch (76 mm) spacing, approximately half of the incident energy was transmitted through the screen. **Fig. 2** shows the results of these tests.

Fig. 3 shows *the results of the second* set of tests. With spacings up to 2¼ inches (57mm), most of the transmitted signal power was reflected into the receiving horn. With 3-inch (76mm) spacing, half the power was reflected and half was transmitted through the screen. This correlates very well with the first set of experiments. For spacings of 3-3/4 inch (95mm) or more, nearly all of the incident energy passed through the screen.

data analysis

The data shows clearly that the screen is a good reflector for element spacings less than 3 inches (76mm) at 2304 MHz, and a poor reflector for spacings greater than 3 inches (76mm). The space between the 7/16-inch (11mm) OD aluminum tubes for a 3-inch (76mm) center-to-center spacing is 2.5625 inches (65mm). Recognize that a free-space half-wavelength at 2304 MHz is very nearly the same value, or 2.563 inches. Further thought suggests that the adjacent elements may operate on much the same principle as a transverse magnetic (TM) mode waveguide.[5] To test this idea, note that the cutoff frequency in such a case depends on the dimensions of the waveguide in the following manner:

$$\lambda_c = \frac{2ab}{\sqrt{a^2 + b^2}}$$

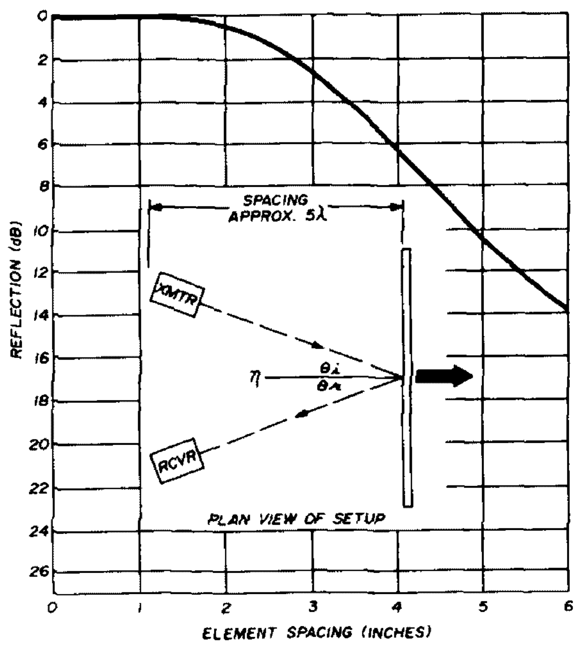


fig. 3. Reflection characteristic at 2304 MHz as a function of reflector element spacing.

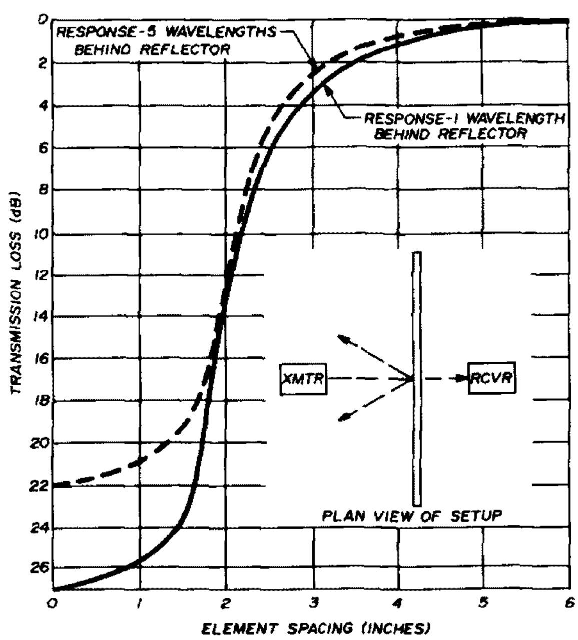


fig. 2. Transmission characteristic at 2304 MHz as a function of reflector element spacing.

This equation can be illustrated geometrically as shown in fig. 4. Note that as the *a* dimension is increased to infinity, which is the case of parallel reflector elements, then $\lambda c/2$ approaches *b*. This is exactly the result that the experiments led to. In other words, by definition, the cutoff frequency of a curtain of parallel reflector elements occurs when the space *between* adjacent elements is one-half wavelength.

This interesting result also implies that the diameter of the reflectors has little to do with the cutoff frequency, since it is the open space between the reflector elements that is significant. Therefore, the analysis also applies to tubes of different diameters as well as to reflector elements having different cross-sectional shapes.

The experiments also show that at the cutoff spacing, the loss in terms of percentage of power transmitted is 3 dB. This can be interpreted to indicate that the illumination efficiency of a parabolic reflector with element spacing corresponding to the cutoff frequency would be 50%. For example, assume a dish diameter of 12 feet (3.7 meters) operating at 2304 MHz; its gain, if it was well designed in all respects, is given by[6]

$$\text{Gain} = 0.59\left(\frac{\pi D}{\lambda}\right)^2 = 0.59\left(\frac{\pi 3.7}{0.13}\right)^2$$
$$= 4703 \text{ or } 36.7 \text{ dB}$$

where D is the diameter of the reflector, λ is the operating wavelength and the coefficient 0.59 accounts for the illumination efficiency of a well designed dish. If the element spacing was the critical value, however, allowing half the incident power to pass through the elements, the dish gain would be 3 dB less, or 33.6 dB. The overall efficiency of such a dish would be 29% instead of 59%.

It is interesting to note that, even with losses of this magnitude, the ERP of this dish would be over 10 kW when fed with 5 watts. If the element spacing was reduced to 2 inches (51mm) however, the ERP would increase to over 20 kW at 2304 MHz.

economical 432-MHz design

On the basis of these experimental and analytical results, a parabolic reflector can be built for 432 MHz having only 1 dB reflector loss if the elements are spaced 12 inches (30.5cm) apart. In this case, only 13 reflector elements are needed for a 12-foot (3.7-meter) dish.

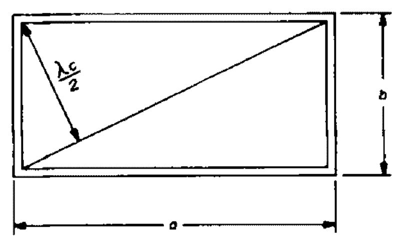


fig. 4. Geometry of the transverse magnetic mode (TM_{11}) in rectangular waveguide (see text).

The design would be very lightweight and would have relatively low wind-cross section. Compared to the ideal dish made of solid sheet metal, this dish would have an illumination efficiency of 47% instead of 59%, and a gain of 21 dB instead of 22 dB.

1296-MHz dish design

It is interesting that 7-foot (2.1-meter) parabolic receiving antennas manufactured for TV use employ 4-inch (10.2 cm) spacing. This makes them useful at 1296 MHz without the addition of hardware cloth or other metal screening if you are willing to accept approximately 1.0 dB reduction in gain compared to a solid metal design. The gain of such a dish is 26 dB at 1296 MHz. If the loss of a dB makes you nervous, the dish diameter can be increased by 12% to compensate for the loss. Thus, a 12-foot (3.6-meter) dish should be made 13 feet (3.96 meters) in diameter, while

the 7-foot (2.1-meter) dish diameter should be increased to nearly 8 feet (2.4 meters) to *compensate for 1 dB of* reflector transmission loss.

It should be recognized that the parallel element dish is sensitive to polarization and will not work well with dual or circularly polarized feed systems such as may be desirable for EME or satellite communications. However, it is entirely feasible to use a second set of reflectors at right-angles to those already described to eliminate polarization sensitivity. Such a dish, if designed for 432 MHz, for example, would have a surface resembling a 12x12-inch (30x30cm) mesh if made of 7/16 inch (11 mm) diameter aluminum tubing.

on the air tests

The 12-foot (3.6-meter) parabolic reflector I use was put into service in 1970. Initially, it was not covered with screening. Tests performed before and after adding expanded aluminum screening both on 432 MHz and 1296 MHz did not indicate any difference in performance or gain.

More recently, comparative tests were run with and without screening on 2304 MHz, and a small reduction in gain was noticed. The radiation patterns on

table 1. Center-to-center spacing of parabolic reflector elements for 1.0 dB loss of gain.

reflector tubing diameter	frequency (MHz) 432	1296	2304
7/16" (11mm)	12.000" (305.0mm)	3.855" (98.0mm)	2.168" (55.0mm)
1/2" (13mm)	12.063" (306.5mm)	9.917" (99.5mm)	2.230" (56.5mm)
5/8" (16mm)	12.125" (308.0mm)	4.042" (102.5mm)	2.355" (60.0mm)
3/4" (19mm)	12.375" (314.5mm)	4.167" (106.0mm)	2.480" (63.0mm)

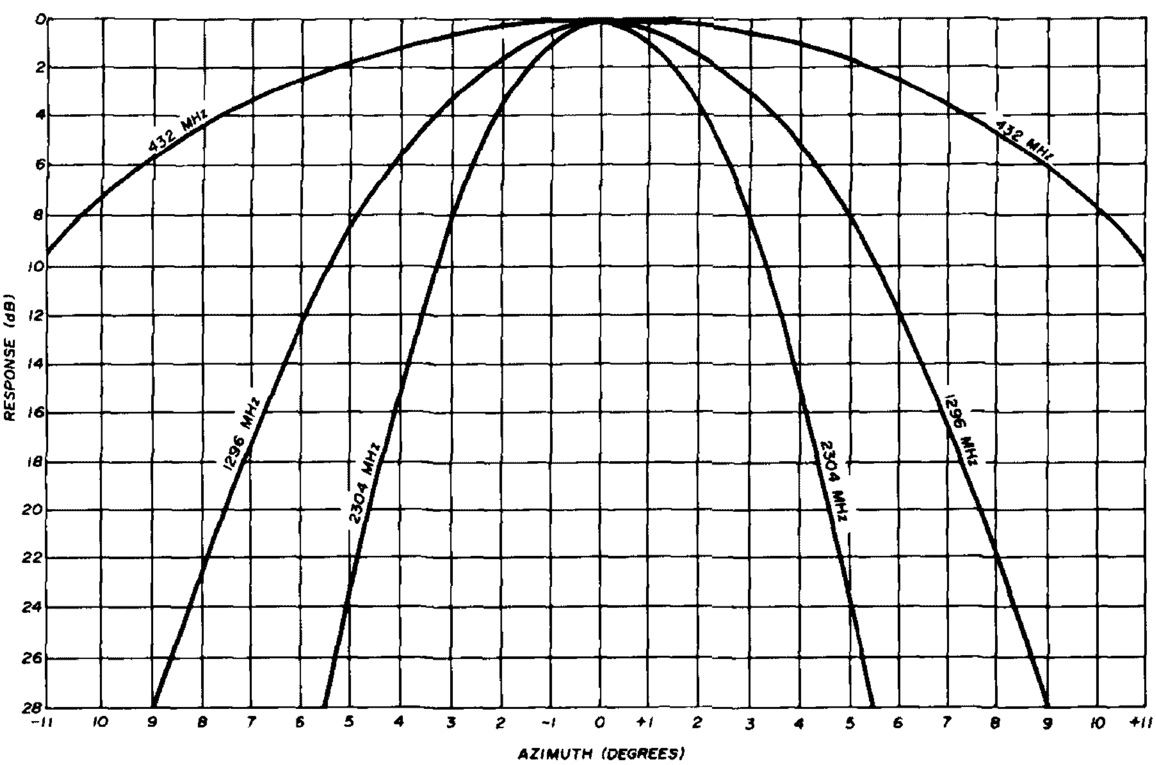


fig. 5. Azimuth patterns of WA9HUV's 12-foot (3.6-meter) parabolic reflector at 432, 1296 and 2304 MHz (photograph of dish is shown at beginning of article).

all three bands did not change however. Fig. 5 shows the patterns of the 12-foot dish without screening for 432, 1296 and 2304 MHz. These patterns were taken with the aid of battery-operated signal sources located approximately 0.2 mile (0.32 km) from the dish, and elevated above ground to the same height as the center of the dish. The pattern test conditions were fairly ideal since the line of sight path was unobstructed, giving essentially free-space conditions.

summary

Probably the most simple and effective parabolic reflector design is the one described in this article. The reflecting elements should be oriented parallel to the direction of polarization of the exciting wave. Thus, for horizontal polarization which is in most general use for amateur purposes, the elements of the parabola should be oriented horizontally. The space between elements should be less than a half wavelength at the highest operating frequency. To reduce the loss through the reflector to 1.0 dB, the space between reflectors should be approximately 0.439λ. Table 1 shows center-to-center spacing for various sizes of tubing for frequencies of 432, 1296 and 2304 MHz.

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ham radio

how to design shunt-feed systems for grounded vertical radiators

A graphical design system for using your tower as a shunt-fed vertical antenna

John R. True, W4QQ, 10322 Georgetown Pike, Great Falls, Virginia 22066

Vertical antennas have several advantages over horizontal dipoles on the lower amateur bands, especially in those cases where the dipole cannot be raised at least one-half wavelength above ground. A recent article showed how to use a 54-foot (16.5m) tower, top loaded with a quad or Yagi, as a grounded vertical radiator on 40 and 80 meters.[1] However, to properly design the shunt-feed matching system for these two lower bands required the use of a good quality impedance bridge. Once the complex input impedance had been determined, a graphical method was used to calculate the components required to match that impedance to a 50-ohm transmission line.[2]

This antenna system generated a great deal of interest, but since few amateurs have access to an RX impedance bridge, they were unable to use this technique to adapt their own towers for use on the lower amateur bands. For this reason I decided to make a series of measurements which would be used to generate a set of graphs which would simplify the design of shunt-fed vertical radiators. These graphs are presented in this article.

First, a series of antenna tests were conducted by scale modeling to determine the electrical height of towers which are capacitance loaded by a typical Yagi beam or cubical quad. Further tests were conducted to determine how long the gamma-type shunt feed rod had to be to permit the use of a practical L-network for matching to 50-ohm coaxial cable.

All tests were made with an aluminum-tubing gamma rod about 1 inch (25mm) outside diameter, spaced 10 ±2 inches (20.3 to 30.5cm) from one leg of the tower. This size was chosen for maximum physical and electrical stability as well as high conductivity. The resultant design curves show the electrical height of the tower, required gamma rod length, and series capacitance, C_s, required to cancel the inductive reactance of the gamma rod. The parallel matching capacitance, C_p, is also given (fig. 1). The series and parallel capacitors should both be air variables so the matching system can be adjusted to provide as low vswr as possible.

using the curves

The graph of **fig. 2** shows the relationship of physical height to electrical height of a thin wire (calculated from

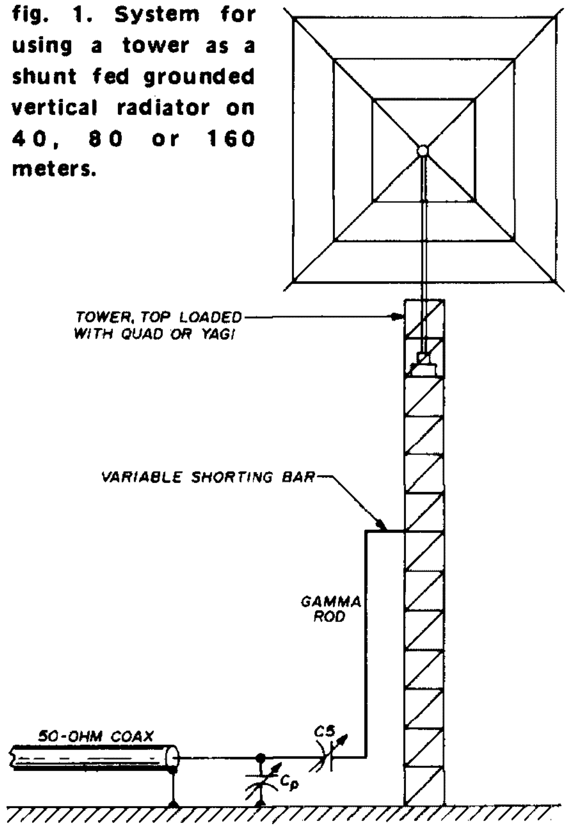
$234/f_{MHz}$), measured electrical height of a 1½-inch (38mm) diameter conductor (which coincides very closely with the predicted electrical height of a thin conductor), and a tower 1 to 2 feet (30 to 61cm) in cross section, top loaded with a Yagi or cubical quad. If you wish to use your present tower as a vertical antenna for the lower bands, you can determine its electrical height from the data presented in fig. 2.

The data in fig. 3 is for use on the amateur 7-MHz band and shows the length of the gamma rod and required series capacitance for towers up to 90 feet (27 meters) high (about 3/4 wavelength on 40 meters). Towers which are taller than this will produce a large lobe of high-angle radiation that reduces the radiation at lower vertical angles. Some shorter towers may actually be shorter, physically, than the recommended gamma rod; in that case more parallel capacitance will be required to match the system to 50 ohms. Fig. 4 shows the same type of data for the 80-meter band (towers higher than 180 feet [54 meters] exhibit the large, high-angle lobe).

The data in fig. 5 is for use on the 160-meter band. Note that a tower which has an electrical height of 90 feet (27 meters) requires a gamma rod which is 60 feet (18 meters) long. Since a 43-foot (12m) tower with a Yagi represents an electrical height near 90 feet, a 60-foot gamma rod is obviously an impossibility. The use of a shorter gamma rod and more parallel capacitance *may* provide a match to 50 ohms, but in this case an rf bridge and graphical solution will save a lot of time.[3]

Note that for towers with electrical heights near 53 and 70 feet (16.2 and 21.3 meters), a gamma rod approximately 20 feet (6.1 meters) long will provide operation on both 80 *and* 40 meters (the rod is about a quarter-wavelength long on 80 meters, one-half wavelength long on 40). For operation on both 80 and 160 meters, a similar coincidence occurs for towers which are electrically near 110 and 135 feet (35.5 and 41.1 meters) high. In this case a gamma rod approximately 40 feet (12.2 meters) long will provide operation on both bands. In either of these dual-band systems adjustments of the parallel tuning capacitor, C_p, will compensate for differences from the specified gamma rod length.

fig. 1. System for using a tower as a shunt fed grounded vertical radiator on 40, 80 or 160 meters.



The electrical height of towers higher than 120 feet (36 meters) can be extrapolated by adding about 35 feet (10 meters) for a three-element 20-meter Yagi with a quarter-wavelength boom (about 16 feet or 5 meters); add about 45 feet (13 meters) of electrical height for a multielement beam such as the Hy-Gain TH6DXX. A two-element 40-meter beam adds 50 to 60 feet (15 to 18 meters). Although cubical quads add about 25 feet (7.6 meters), multi-element quad designs add little more because the elements are well away from the top of the tower and insulated from it.

matching capacitors

Since the reactance of the series capacitor, C_s, is quite large except in those cases where the tower is approximately a quarter-wavelength high, this capacitor should have a breakdown rating of about 1000 volts for transmitters up to about 200 watts output. For transmitter powers of 2000 watts this capacitor should have a breakdown rating of 5000 volts or more.

Where large capacitance values are recommended, it is suggested that at least half be variable with the rest made up with fixed padding. Note that *both* the stator and rotor of the series capacitor must be isolated from ground.

The ideal matching network for this antenna system would use two vacuum-variable capacitors. These capacitors are not seriously affected by humidity or changes in barometric pressure, and

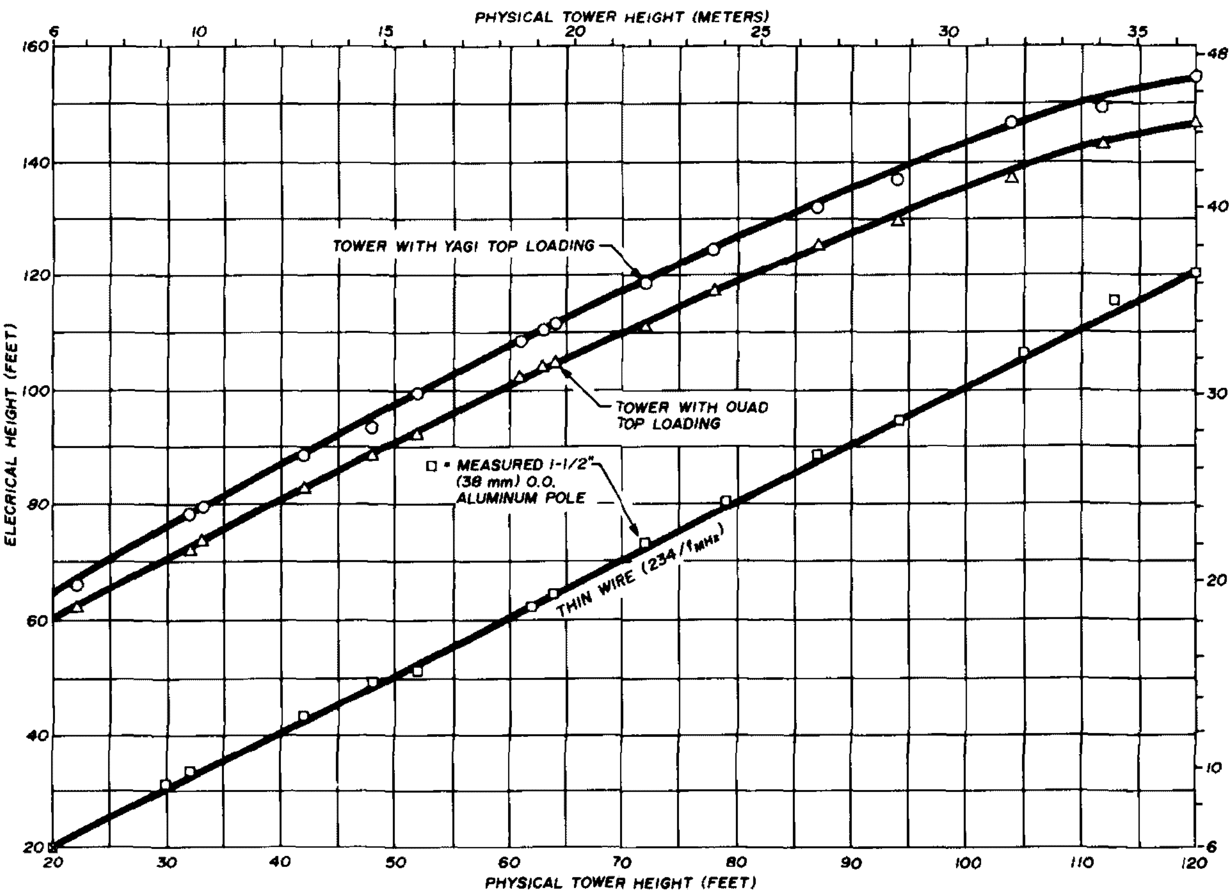


fig. 2. Physical vs electrical height of towers top loaded with Yagi beams or cubical quads.

The parallel matching capacitor, C_p, does not require such a high voltage rating unless excessively high vswr is expected at full power. For a 200-watt transmitter, an old style BC capacitor with 700 to 1000 pF maximum should work nicely. For 2000 watts PEP the parallel capacitor should have a rating of 1500 volts minimum with current-carrying capacity of seven amperes.

they can be connected to small geared motors so they can be controlled remotely from the operating position. A 300-pF vacuum variable rated at 7500 volts, and a 1000-pF vacuum variable with a 2000 volt rating should handle practically any legal amateur transmitter with low vswr.

A remote-control system that I have used for several years is shown in refer-

ence 1. It's obviously a lot easier to remotely control the matching system from your hamshack than it is to traipse out to the backyard in snow, sleet and rain each time you want to shift your operating frequency.

construction

A typical gamma rod installation is shown in fig. 6. On my vertical antenna the gamma rod is mounted with PVC each side of the PVC pipe, about 1 inch (25mm) in from each end (see fig. 6). Stainless-steel hose clamps are run through the slits in the PVC pipe and around the vertical member.*

If you wish, the same tower may be used on more than one lower-frequency band – simply install gamma rods on more than one leg of the tower. You can use separate capacitance matching systems or remotely controlled vacuum-

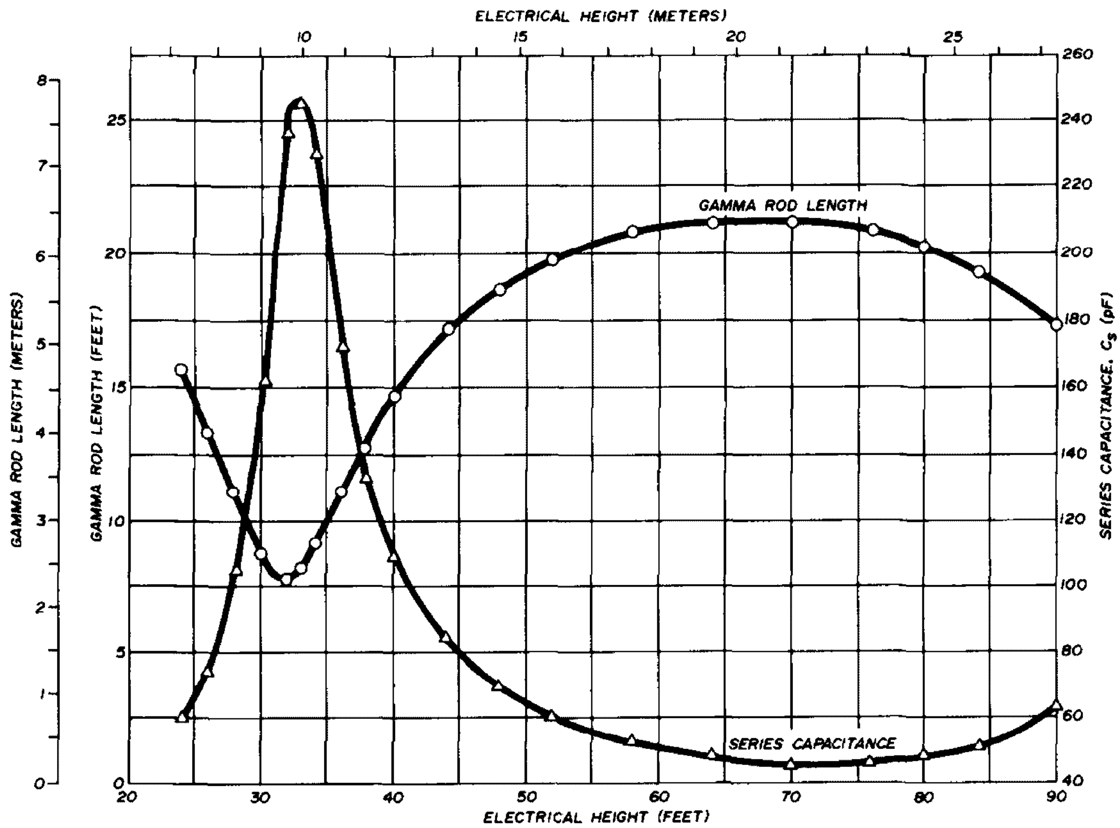


fig. 3. 40-meter vertical. Gamma rod length and series capacitance vs electrical height of tower. Recommended parallel capacitance to match 50-ohm transmission lines is 320 pF (at least 100 pF of which should be variable).

insulators spaced about 10 feet (3 meters) apart. The insulators are made from 1-inch diameter (25mm) PVC water pipe. The movable shorting bar is made from the same material as the gamma rod.

To attach the PVC insulators to the gamma rod, first notch the ends so one end fits around one leg of your tower, the other end around the gamma rod. Then cut half-inch (13mm) long slits on variables, depending upon your operating requirements. A vertical tower antenna system which I use successfully on both 40 and 80 meters is described in reference 1.

*The author has assembled several pages of how-to hints and additional constructional information which is available from him for the cost of printing and mailing. A self-addressed, stamped envelope to the author will bring a summary of contents and cost.

ground requirements

Remember that the vertical element is only one-half of a vertical antenna system – the vertical element must operate against a good ground plane or the ground losses will be so high that the antenna performs poorly. The so-called ideal ground system consists of 120 equally-spaced, quarter-wavelength radials, but even such an elaborate ground plane as this still introduces about 2 ohms of series loss resistance into the total radiation resistance. Since short vertical antennas are characterized by relatively low radiation resistance, ground resistance loss is higher, proportionately, than it is with vertical elements which are quarter-wavelength or more. A complete discussion of ground system requirements is contained in reference 4. For more information on short vertical antennas, consult the excellent series of articles by W2FMI.[5-8]

The tower which you use to support your higher frequency antennas can easily be used as a practical antenna system for 40, 80 and 160 meters. The graphs presented here will help you to design the necessary shunt-matching system, but note that since conditions vary from one location to another, some adjustments will be necessary to obtain a low vswr. However, with an swr bridge installed near the base of the vertical (very short leads), alternately adjust the series and parallel tuning capacitors until the reflected power approaches zero. If the amount of parallel capacitance for low vswr seems excessive, make the gamma rod slightly longer.

The setting of the series capacitor is rather critical because reactance changes

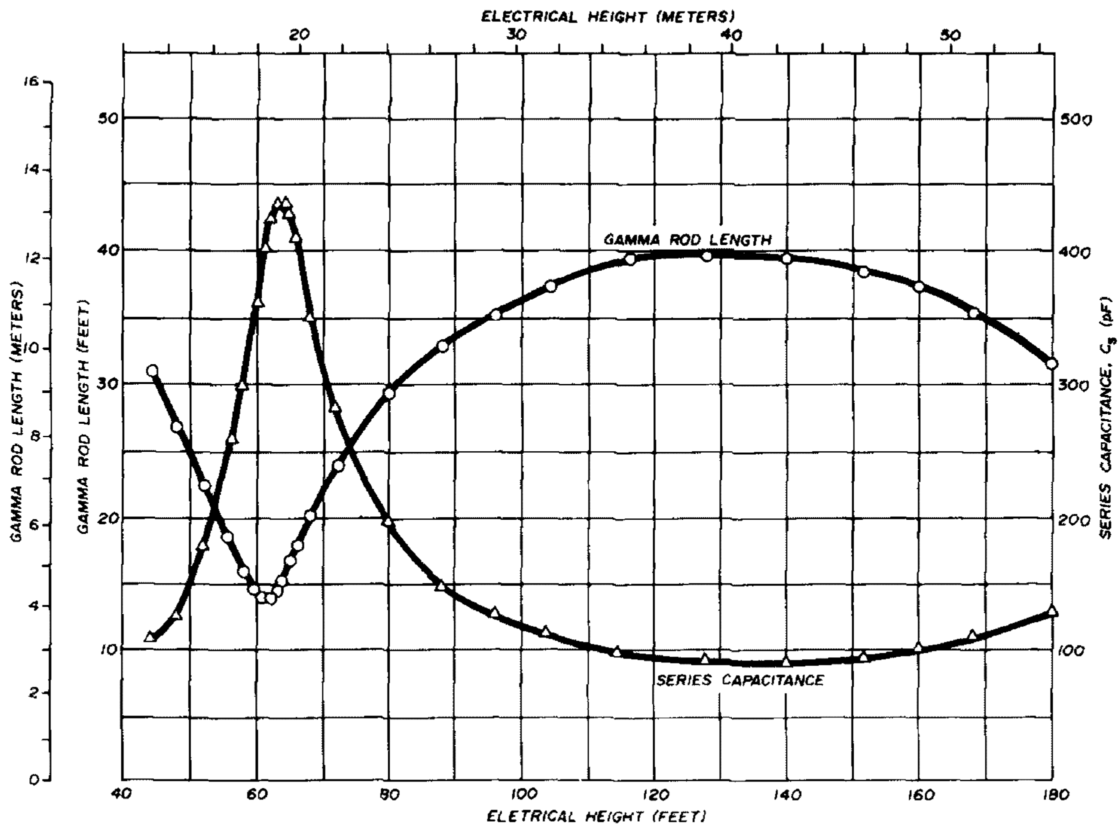


fig. 4. 80-meter vertical. Gamma rod length and series capacitance vs electrical height of tower. Recommended parallel capacitance to match 50-ohm transmission lines is 650 pF (at least half should be variable).

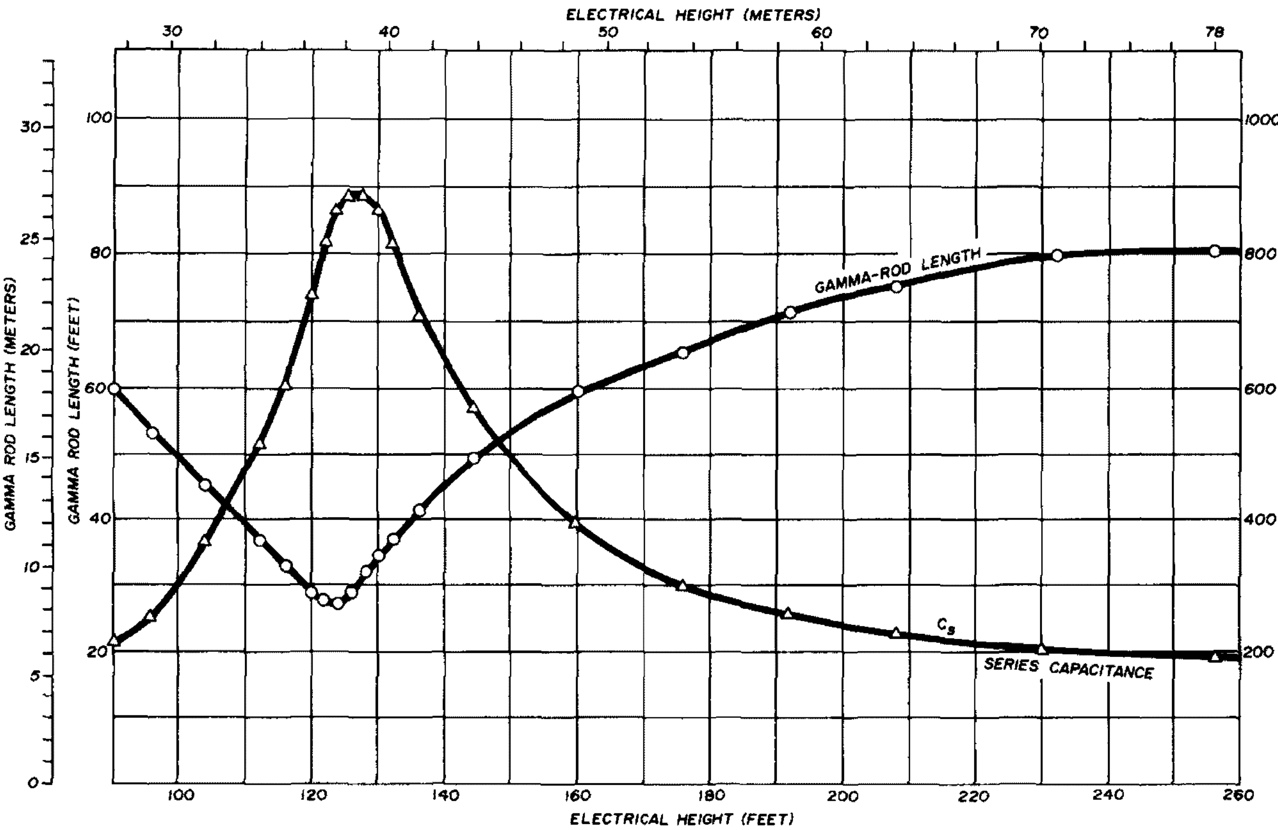


fig. 5. 160-meter vertical. Gamma rod length and series capacitance vs electrical height of tower. Parallel capacitance required to match 50-ohm transmission lines is approximately 1300 pF.

sharply near zero so it may take several tries before you can get the capacitor set exactly right. However, with a good ground system, the shunt-fed grounded tower can provide a very efficient antenna system for relatively little cost.

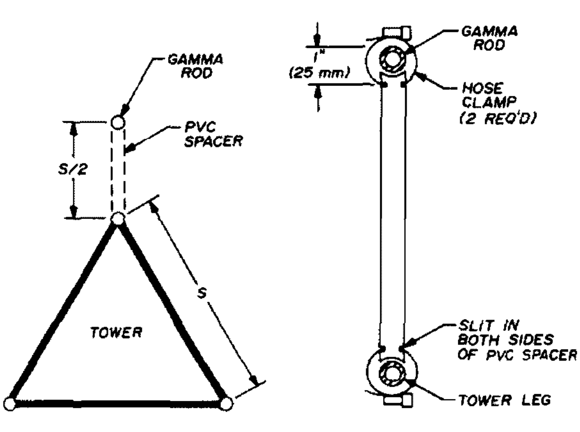
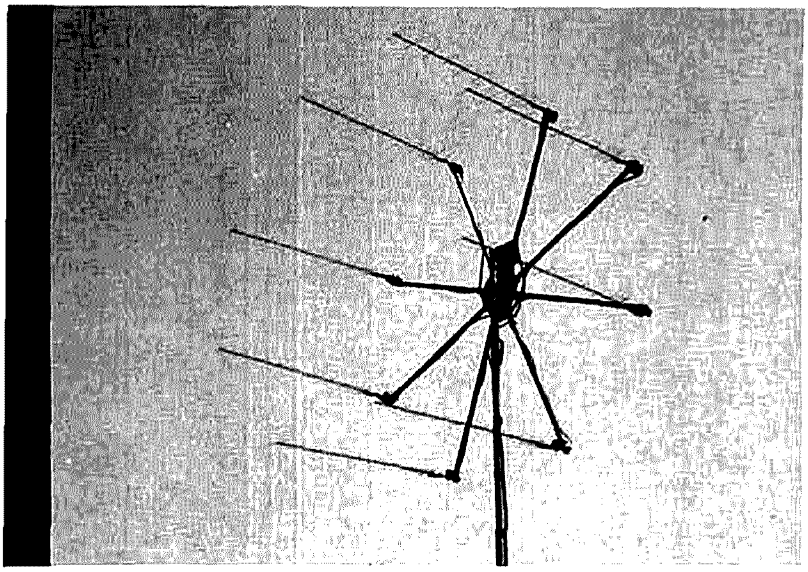


fig. 6. Construction of the shunt feed system for grounded vertical radiators. The spacers are made from PVC water pipe.

references

1. John R. True, W4OQ, "Vertical Antenna System," *ham radio*, April, 1973, page 16, and May, 1973, page 56.
2. I.L. McNally, W1NCK, and Henry S. Keen, W2CTK, "Graphical Solution of Impedance-Matching Problems," *ham radio*, December, 1969, page 26.
3. John R. True, W4OQ, "Measurement Techniques for Antennas and Transmission Lines," *ham radio*, May 1974, page 36.
4. Robert E. Leo, W7LR, "Vertical Antenna Ground Systems," *ham radio*, May, 1974, page 30.
5. Jerry Sevick, W2FMI, "The W2FMI 20-Meter Vertical Beam," *QST*, June, 1972, page 14.
6. Jerry Sevick, W2FMI, "The W2FMI Ground-Mounted Short Vertical," *QST*, March, 1973, page 13.
7. Jerry Sevick, W2FMI, "A High Performance 20-, 40- and 80-Meter Vertical System, *QST*, December, 1973, page 30.
8. Jerry Sevick, W2FMI, "The Constant-Impedance Trap Vertical, *QST*, March, 1974, page 29.

ham radio



high-gain 1296-MHz Yagi Array

Construction of a light-weight high performance 104-element antenna for 1296 MHz

Paul F. Magee, W3AED, Route 2, Box 432, Berlin, Maryland

The high-gain 1296-MHz array described in this article, which I call the "blowtorch array" and which consists of eight 13-element Yagis arranged in a circle, is light weight, presents little wind resistance, can be rotated with a small TV rotator, and provides gain equivalent to that of a 5-foot (1.5 meter) dish. The Yagis, based on a design by W2CQH,[1] use slightly shorter directors than he specified and are designed for use with low-loss 75-ohm CATV coaxial feedline.

yagi construction

The parasitic elements of the Yagis are lengths of number-14 (1.6mm) OD copperweld wire, soft-soldered to a boom made of ¼-inch (6.5mm) thick-wall brass tubing, 36 inches (91.4cm) long as shown in fig. 1. The copperweld elements provide both physical strength and high electrical conductivity, both of which are required for effective operation.

Before cutting the directors, make up a simple template as shown in fig. 2 so that each of the directors is precisely the same length. This is extremely im-

*Any specific information desired regarding this array can be obtained by enclosing a self-addressed stamped envelope with inquiry to Paul F. Magee, W3AED, R2 Box 432, Berlin, Maryland 21811.

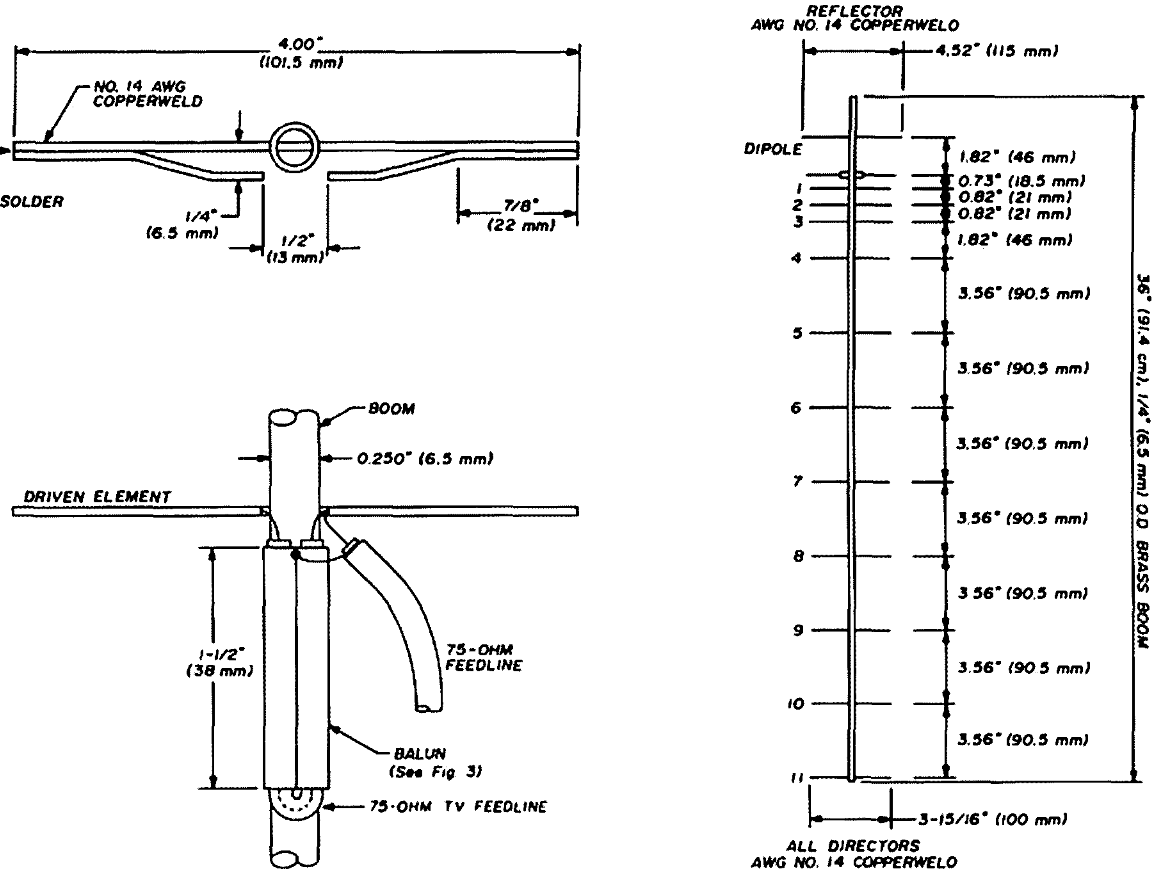
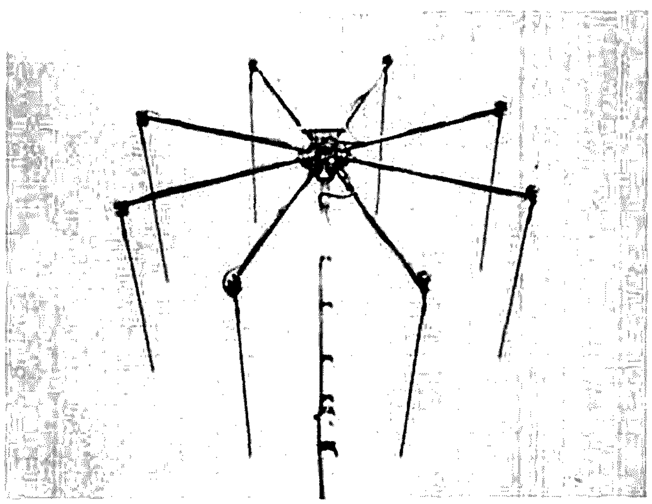


fig. 1. Layout of the 13-element Yagis used in the 1296-MHz blowtorch array. Details of 75-ohm feed system are shown in fig. 3.

portant because very slight differences in length will adversely affect the performance of the antenna.

The driven element is a delta-matched dipole of number-14 copperweld which is soldered to the boom. The delta-matching system consists of two pieces of number-14 soft copper wire soldered to the dipole — this is fed by a balun made from two pieces of ¼-inch (6.5mm) OD copper tubing, 1½ inch (38mm) long, which are soldered on top of the boom as shown in fig. 3. Place the forward end of these two tubes as close as possible to the driven element to minimize lead length to the delta match.

To hold the two balun tubes in the proper position while you are soldering them, bend a small piece of sheet aluminum into a vee which can be held loosely in a vise. Note the short number-14 stud which is soldered between the quarter-wavelength balun tubes — this provides a convenient place to solder the shield of the 75-ohm coaxial phasing



Rear view of the 1296-MHz blowtorch array showing the waterproof enclosure which houses the quarter-wavelength matching transformers.

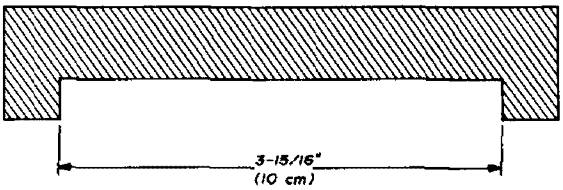


fig. 2. Use of a template is recommended to assure that each of the directors is precisely the same length. For best results, cut each of the directors slightly longer than necessary and file off the ends until each element just passes through the guage.

line (be sure the phasing lines are connected to the same side of the delta match on all eight Yagis).

After the two copper balun tubes and all elements have been soldered to the boom, strip off the outer jacket and shield from a section of 75-ohm TV feedling (which has a larger diameter than RG-59/U). Fold the balun wire into a U and insert it into the balun tubes. Connect the center conductors to each side of the delta match.

The antenna mount consists of a center hub of 1/8-inch (3mm) thick galvanized steel, 8 inches (20cm) in diameter, to which is attached a spider-shaped arrangement of 3/4-inch (19mm) aluminum tubes, each 31 inches (79cm) long. Short tabs on the top and bottom of the center hub provide space for the U-bolts which are used to attach the array to a mast.

The aluminum tubes are attached to the hub with 10-32 screws and strengthening members made from 3-inch (76mm) lengths of 1-inch (25mm) square aluminum channel (see fig. 4). The Yagis are attached to the ends of these tubes with special clamps which

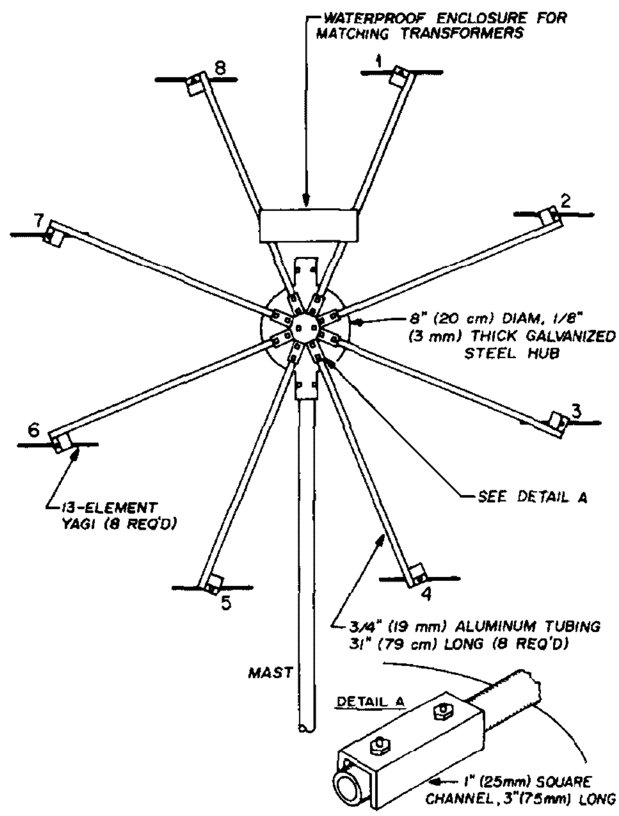


fig. 4. Construction of the antenna mount showing the center hub and eight aluminum antenna-support masts. Main mast is placed in front of the center hub and attached to it with U-bolts.

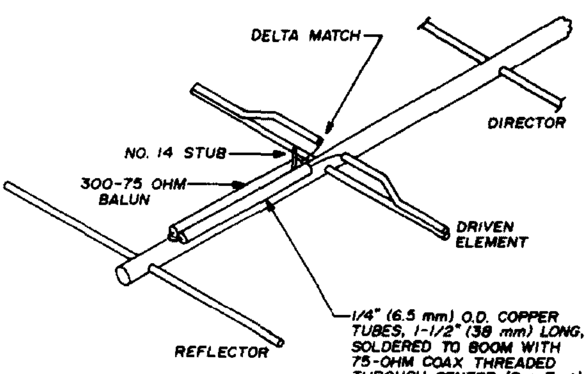


fig. 3. Construction of the 300- to 75-ohm balun used on each of the Yagis. The shield of the coaxial feedline is connected to the short stud.

allow the antennas to be oriented parallel to one another.

The Yagi clamps, shown in fig. 5, are made from 2-inch (51mm) lengths of 1-inch (25mm) square aluminum channel which is attached to the aluminum tubes with small U-bolts made from ¼-inch (6mm) threaded steel stock. Two small notched aluminum blocks are used to hold the antenna boom to the channel (cut the notch

slightly smaller than the diameter of the boom so the antennas are held very rigidly).

Be sure to use brass nuts on the steel U-bolts. The combination of aluminum clamps and brass hardware doesn't seem to cause any problem as my array shows no ill effects although it has been exposed to the salt air at my station for more than two years.

matching harness

The matching system for the blowtorch array is made up of a system of 50-ohm quarter-wave matching transformers and 75-ohm coaxial feedlines 16 half-wavelengths long (48 inches or 122cm). A schematic diagram of the system is shown in fig. 6. Since the 75-ohm matching lines are an integral number of half-wavelengths long, the 75-ohm input impedance of each of the antennas is repeated at the opposite end. Since two 75-ohm feedlines are connected in parallel at points A, C, E and G, the impedance at each of these points is 37.5 ohms. This is transformed to 75 ohms by a 50-ohm quarter-wave transformer. These transformers are connected in parallel at points B and F, so the impedance at these points is 37.5 ohms. This is transformed to 75 ohms with another 50-ohm quarter-wave transformer (B-D and F-D).

When connected in parallel at point D and transformed with another quarter-wave transformer, the input impedance provides a close match to the 75-ohm feedline. Note that the 8-wavelength matching lines are not connected to adjacent antennas, but to antennas on opposite sides of the array.

The quarter-wavelength matching transformers are soldered to a 3x5-inch (8x13cm) copper sheet as shown in the photograph of fig. 7. The outer conductor of each quarter-wave transformer consists of a 3/16-inch (5mm) OD copper tube, 1½ inch (38mm) long. Short lengths of RG-58/U coax (strip-

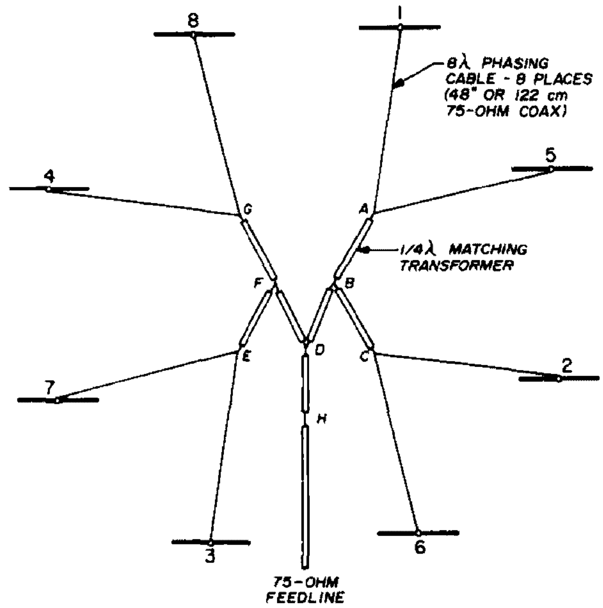


fig. 6. Schematic diagram of the phasing system used with the blowtorch array. Construction of the quarter-wavelength matching transformers is shown in fig. 7.

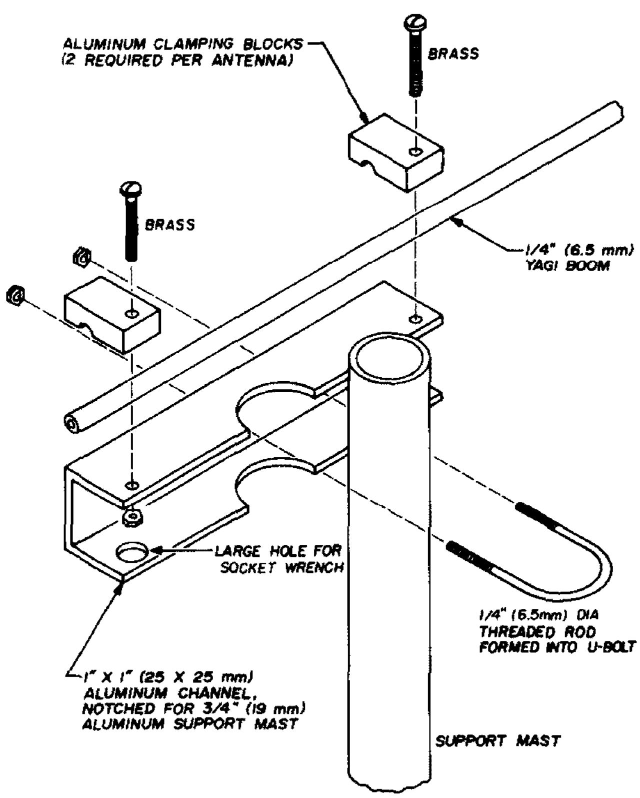


fig. 5. The Yagis are attached to the aluminum support masts with clamps (8 required) which allow all of the beams to be aligned in the same plane.

ped of the outer shield and jacket) are threaded through these tubes. Short stubs of number-14 wire, about 3/8 inch (10mm) high, at the feedline end of each of the transformers provides a convenient point for connecting the outer shield of the 75-ohm feedlines. The completed transformer is mounted in a waterproof enclosure which is installed at the center of the array.

be checked individually before they are installed on the large array. First check the vswr and make adjustments as necessary to the delta match for a vswr of 1.2:1 or less. Then check the lobes on each side of the main pattern. If they are not at least 9 dB down, the antenna is not working correctly. This can be checked by connecting a low-power 1296-MHz signal source to the antenna

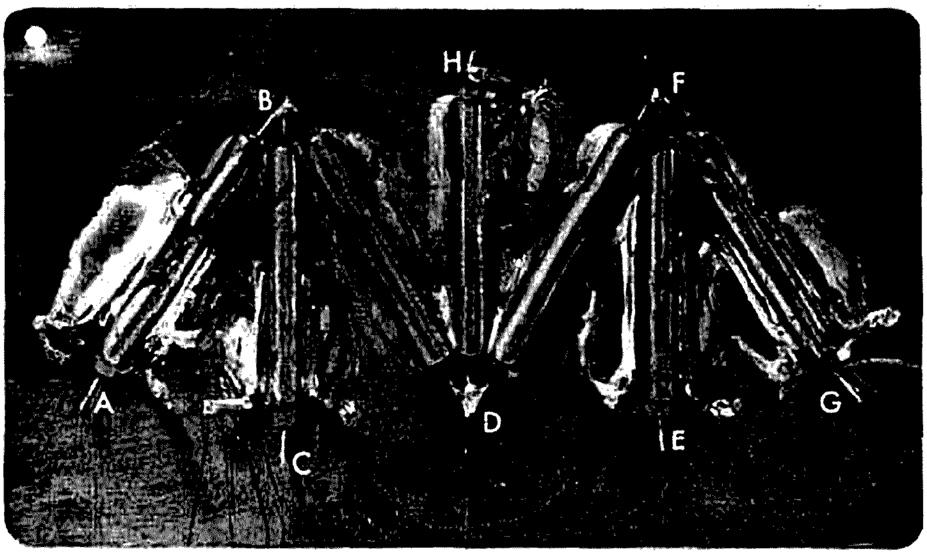


fig. 7. Quarter-wavelength matching transformers are made from 3/16-inch (5mm) OD copper tubing. Letters coincide with the points noted in fig. 6.

When soldering the tubes to the copper sheet make sure they are soldered from one end to the other. You can use a small torch for this, if you wish, but I suggest a husky 200-watt soldering iron. When the transformer assembly has cooled off, the rosin flux can be removed with lighter fluid. (Acid flux can also be used but thoroughly clean off any residue with hot water, baking soda and a tooth brush before you install the sections of RG-58/U.)

testing

Each of the 13-element Yagis should

and measuring the relative field strength about 50 feet (15 meters) in front of the antenna.

When the blowtorch array is completely assembled check the field strength about five feet (1.5 meter) in front of each Yagi; if the phasing system is working properly the field strength in front of each of the antennas should be the same.

reference

1. Reed E. Fisher, W2CQH, "A Successful 1296-MHz Yagi," *ham radio*, May, 1972, page 24.

ham radio

measuring complex impedance with an swr bridge

Accurate impedance measurements can be made over a wide frequency range using a vswr bridge and the simple second-data method described here

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Knowledge of the resistive and reactive components of a load is very useful in antenna work, but few amateurs, unfortunately, have the necessary equipment for making the necessary measurements. The simple impedance bridges found in many amateur workshops will measure resistive loads from about 5 to 400 ohms, and if the load is reactive, a true null will not be obtained. For low impedances the readings of these simple instruments approximate the resistive component while for high impedances the readings approximate the total impedance. In between, the readings are a poor approximation to either.

The RX meter, which can accurately evaluate both the resistive and reactive components of a load, on the other hand, is considerably more complex than the simple rf impedance bridge, and is somewhat more difficult to build and operate, especially above about 30 MHz. A practical amateur RX meter described recently has relatively good sensitivity and a moderate impedance range and doesn't require much drive power, but requires a separate inductance assembly for each band of frequencies.[1] Other homebuilt instruments for measuring complex impedances are described in references 2, 3 and 4.

second-data impedance measurement

Because of these problems and others most amateurs try to avoid the measurement of unknown complex impedance. Fortunately, however, there is an alternative: the second-data method. This method permits relatively accurate measurement of complex impedance with a vswr bridge and an external resistor. Since most amateurs have vswr bridges this is a technique that should not be overlooked. Since vswr bridges which work to 600 MHz and above are readily available,[5] this method has the additional advantage of extending the frequency range.

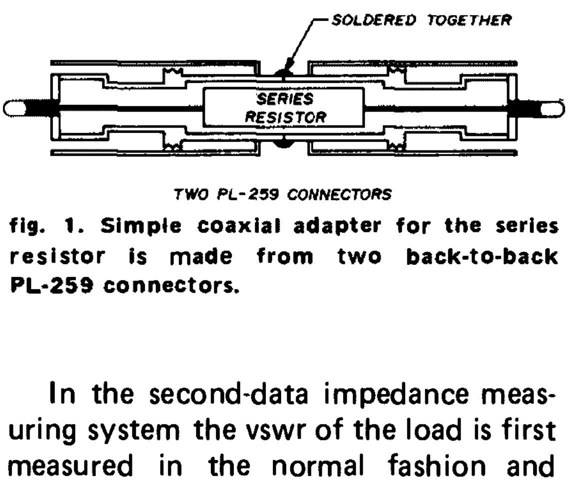

fig. 1. Simple coaxial adapter for the series resistor is made from two back-to-back PL-259 connectors.

In the second-data impedance measuring system the vswr of the load is first measured in the normal fashion and then the vswr is measured with a resistor in series with the load. These two sets of data are adequate to specify the resistive and reactive components of the load. You merely look up the two vswr measurements in a table such as that shown in **table 1** and read off the resistive and reactive components. If you wish, the resistive and reactive values may also be plotted on a Smith chart, as will be discussed later.

For measurements up to 20 MHz or so, the series resistor may be soldered in series with the load or clipped in. For higher frequency measurements or when a coaxial line is used, the series resistor should be installed as the center conductor of an appropriate adapter as shown in **fig. 1**. The vswr is then measured with and without this adapter in the line.

For best results measurements should be made with a vswr bridge that has good linearity characteristics. The resistance in series with the diode detectors within the bridge should have as high a value as possible, at least 10 kilohms. Check your vswr bridge for no return power (vswr = 1:1) with a good quality dummy load. You can check its accuracy at other standing-wave ratios by using standard composition resistors (with short leads) as loads. The vswr of the load is given by

$$\text{vswr} = \frac{R}{Z_o} \text{ or } \frac{Z_o}{R}$$

whichever is greater than 1. Loads of either 150 ohms or 16.67 ohms, for example, will result in a vswr of 3.0:1 in a 50-ohm system.

The vswr to series complex impedance conversion information shown in **table 1** for 50-ohm systems is based on the use of a 27-ohm series resistor. The resistive and reactive values were calculated with an error of less than one percent and rounded off to the nearest ohm.* To illustrate the use of this table, assume that an unknown load exhibits a vswr of 2.4:1. When the 27-ohm resistor is placed in series with the load the measured vswr is 1.8:1. From **table 1**, the resistive component of the load is 33 ohms; the reactive component, 32 ohms.

*More detailed, computer-generated tables for both 50- and 75-ohm systems are available from the author for $4.00, postpaid. The tables are calculated for 27-ohm series resistance with 50-ohm systems and 39-ohm series resistance with 75-ohm systems.

table 1. *Vswr to series complex impedance conversion for 50-ohm systems using the second-data method of measurement (series resistor = 27 ohms).*

vswr of load	vswr with series resistor	resistance (ohms)	reactance (ohms)	vswr of load	vswr with series resistor	resistance (ohms)	reactance (ohms)
1.0	1.53	51	0	3.0	1.0	17	0
				3.0	1.2	17	5
1.2	1.3	43	0	3.0	1.4	19	15
1.2	1.4	44	5	3.0	1.6	21	23
1.2	1.5	48	8	3.0	2.0	28	37
1.2	1.6	53	9	3.0	2.2	33	44
1.2	1.7	59	6	3.0	2.6	47	56
1.2	1.8	61	3	3.0	3.0	71	67
				3.0	3.4	122	57
1.3	1.2	39	0	3.0	3.6	152	13
1.3	1.4	42	8				
1.3	1.6	51	13	4.0	1.25	13	0
1.3	1.8	64	8	4.0	1.5	14	16
1.3	1.9	66	4	4.0	1.75	16	25
				4.0	2.0	19	34
1.4	1.2	36	0	4.0	2.25	22	42
1.4	1.3	38	6	4.0	2.5	27	50
1.4	1.4	40	11	4.0	3.0	38	65
1.4	1.6	48	17	4.0	3.5	58	81
1.4	1.8	61	16	4.0	4.0	95	95
1.4	2.0	71	4	4.0	4.5	193	45
1.5	1.1	34	0	6.0	1.25	9	0
1.5	1.3	36	9	6.0	2.0	11	28
1.5	1.4	38	13	6.0	3.0	19	56
1.5	1.5	42	17	6.0	4.0	34	83
1.5	1.6	46	19	6.0	4.5	45	98
1.5	1.7	51	21	6.0	5.0	62	114
1.5	1.8	56	21	6.0	5.5	90	133
1.5	1.9	64	19	6.0	6.0	143	148
1.5	2.0	73	11	6.0	6.5	288	71
1.5	2.1	76	4	6.0	6.75	306	11
2.0	1.0	26	0	8.0	1.5	6	0
2.0	1.2	27	9	8.0	2.0	8	25
2.0	1.4	30	17	8.0	2.5	10	39
2.0	1.6	34	24	8.0	3.0	13	51
2.0	1.8	41	31	8.0	4.0	20	74
2.0	2.0	50	36	8.0	5.0	32	98
2.0	2.2	64	38	8.0	6.0	51	126
2.0	2.4	83	33	8.0	7.0	87	161
2.0	2.6	102	5	8.0	8.0	194	200
2.0	2.8	102	5	8.0	8.5	385	94
2.4	1.0	21	0	10.0	1.5	5	0
2.4	1.2	22	9	10.0	2.0	6	23
2.4	1.4	24	17	10.0	3.0	10	48
2.4	1.6	28	25	10.0	4.0	14	68
2.4	1.8	33	32	10.0	5.0	21	89
2.4	2.0	39	38	10.0	7.0	45	136
2.4	2.2	47	44	10.0	8.0	68	166
2.4	2.6	74	51	10.0	9.0	114	207
2.4	2.8	98	43	10.0	10.0	245	252
2.4	3.0	121	11	10.0	10.5	481	117

Since the entire range of possible vswr combinations for a series 27-ohm resistor is given in table 1, if you measure a larger vswr than shown, there is an error in measurement. For example, if you measure a load vswr of 2.0:1, then install the 27-ohm resistor and measure the vswr again, a new vswr measurement of 4.0:1 reveals an error in measurement as it is outside the range of values given.

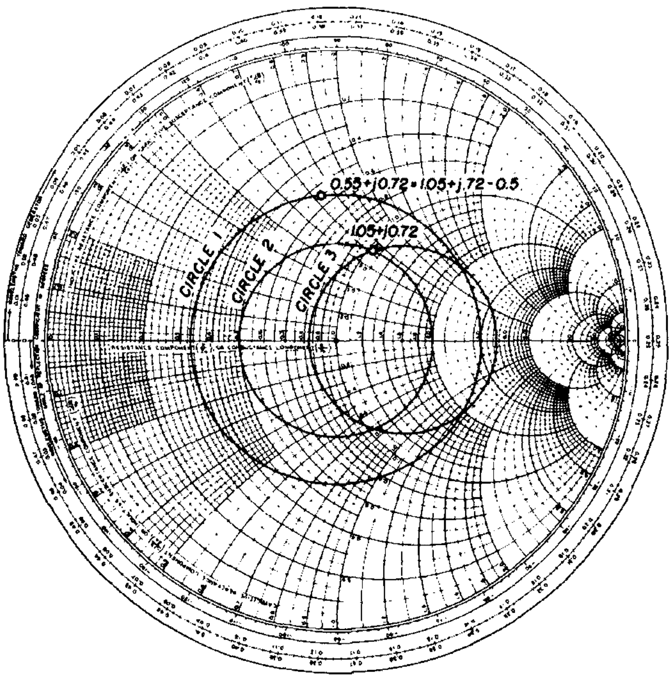

fig. 2. Resistive and reactive impedance values can be plotted on Smith chart using information obtained by the second-data method. Circle 1 represents the vswr of the load, circle 2 represents the vswr of the load with series resistor, and circle 3 shows the vswr of circle 2 displaced by the amount of the resistance (see text).

smith chart computation

If you wish to use the Smith chart to make your impedance calculations, the series resistor does not have to be the same value as that used in computing the chart (27 ohms for 50-ohm systems) – it can be any reasonable value. Assume, for example, that the series resistor is 25 ohms, the measured vswr of the unknown load is 3.0:1, and with the 25-ohm resistor in series with load the vswr is 2.0:1.

To compute the complex impedance of the unknown load you must first plot two circles on the Smith chart as shown in fig. 2. The first circle, circle 1, is a constant vswr circle corresponding to a measured vswr of 3.0:1 and intersects with 3.0 on the real axis. This circle passes through all the resistive and reactive pairs which produce a vswr of 3.0:1, so the impedance of the unknown load must fall somewhere on this circle.

The second circle, circle 2, is another constant vswr circle, this one corresponding to the 2.0:1 vswr measured with the 25-ohm resistor in the line. The final circle, circle 3, which is offset by the amount of resistance in series with the line for the second vswr measurement, represents all the values on circle 1 with $Z_o/25$ (0.5 resistive) added to each point. This circle is easily drawn by adding 0.5 resistive to both points where the circle crosses the real axis (at 0.33 + 0.5 = 0.83 and 3.0 + 0.5 = 3.5), and plotting a circle through these two points.

Circle 3 intersects circle 2 at two points. Both points have the same resistive and reactive values, but the reactive values have opposite signs. To transform the intersection on circle 2 to the actual value of the unknown load, 0.5 ohm resistive must be subtracted from the values on circle 2. In this case the intersection on the upper half of the chart is 1.05 + j0.72. Subtracting 0.5 ohm resistive yields 0.55 + j0.72. This is the original load and should fall on circle 1. If it doesn't, you made an error in your calculations.

Since this is a normalized Smith chart, the plotted values must be multiplied by the characteristic impedance of the system, 50 ohms. This yields values of 27.5 ohms resistive and 36 ohms reactive for the unknown load. You must still determine whether the reactance is capacitive or inductive (negative

or positive). In some cases you may know beforehand whether the reactance is capacitive or inductive (a half-wave dipole operated above resonance, for example, has an input impedance that is inductive) but in those cases where you must determine the sign of the reactance another measurement is required. After the resistive and reactive components of the load have been determined, the vswr is measured again with a capacitive reactance ($X_c \approx X_L$) placed in series with the load. If the resulting vswr is *less* than the original vswr, the reactance of the load is *inductive*. If the resultant vswr is *greater* than the original vswr, the reactance is *capacitive*.

Things are simplified considerably when the impedance measurement can be made directly at the input terminals of the unknown load, but this is not always possible. The usual solution is to take the measurement at the end of a transmission line and use the Smith chart to refer the load back to the antenna feedpoint. Further information on this subject is provided in references 6 and 7.

"How does it load on 160?"

accuracy

The accuracy of this method of impedance measurement, of course, is dependent upon the accuracy with which the vswr measurements are made. Accuracy of the second-data method is also dependent on the magnitude of the measured impedance with errors increasing with increasing impedance. However, accuracy is acceptable up to about 300 ohms, depending on the accuracy of the vswr bridge, and accuracy is quite good around the commonly used system impedances of 50 and 75 ohms.

As an example, consider the case of a quarter-wavelength vertical operating over perfect ground. Its base feedpoint impedance, theoretically, is 36 ohms. If the second-data technique is used with a vswr bridge that reads 10% high, the indicated impedance will be 36.7 ohms resistive and 12.7 ohms reactive; the bridge has introduced a small reactive component. As with any electronic measurements, good results are guaranteed only when the measuring instruments have been carefully checked and calibrated.

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ham radio

electrically-controlled phased array

Phased array for the lower amateur bands uses hybrid couplers for electrical control of the field pattern

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On the lower frequency amateur bands, where a rotary antenna may be too cumbersome for convenience, the answer may be a phased array, with electrical instead of mechanical control of the field pattern. Such a system will also be less vulnerable to damage from ice and high winds.

The phasor, which is the heart of this system, is built around a 3 dB quadrature hybrid. Such a hybrid may be regarded as a four-port circuit element, having two input ports and two output ports (**fig. 1**). When both output ports are terminated by matched loads, a signal injected at either input port will divide equally between the two output ports with a 90°, or quadrature, phase difference between the outputs. Because no signal exists at the remaining input port, the input ports may be seen to be isolated from each other.

Several different designs are available to the experimenter for the construction of such a hybrid.[1] The most compact and convenient form for the lower amateur bands, however, consists of two capacitively coupled lengths of coaxial line, of whichever characteristic impedance has been chosen for the system (**fig. 2**). These lengths of coaxial line will each be an electrical one-eighth wavelength at the frequency of interest, while the coupling capacitors will have a reactance equal to the characteristic impedance of the coaxial lines.

theory of operation

If the output ports of the hybrid, instead of being terminated by a pair of matched loads as in **fig. 1**, are terminated instead with identical lengths of transmission line as shown in **fig. 3**, the following action will take place as a signal is injected into port 1:

1. The injected signal will divide into two equal parts, one going to port 3, while the other part, after undergoing a 90° phase shift, appears at port 4. Both signals then progress down to the ends of their terminating lines and are reflected back to their respective ports, having traveled back and forth along the line for a total distance equal to twice the actual length of the line.

2. The reflected signal returning to port 3 divides into two equal parts, one of which goes directly to port 1, while the other part, after a 90° phase shift, goes to port 2.

3. The reflected signal at port 4 also divides into two equal parts, one of which goes directly to port 2, where it combines with the component which came from port 3. Note that each of these two components has traveled the same distance and undergone a single

phase shift of 90°. The other part of the reflected signal at port 4 undergoes a second 90° phase shift and goes to port 1. Both of the components at port 1 can be seen to have traveled the same distance, but are 180° out of phase with each other, and therefore cancel. As a

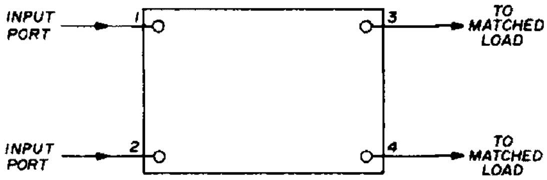


fig. 1. Block diagram of hybrid coupler. This device divides power between two matched loads.

result, no signal is reflected back to port 1 which is not cancelled out, and no input vswr results. It may be seen, therefore, that any equal change in the lengths of the two terminating lines will result in a phase change in the signal emerging from port 2 which is equal to twice the change in line lengths.

With this last conclusion in mind, it becomes apparent that a reactance change at ports 3 and 4 which is equivalent to that produced by a quarter wavelength, or 90° change in line length, will result in a total phase change of 180° in the signal emerging from port 2. Using lumped constants, instead of transmission lines, to terminate ports 3 and 4, the network shown in fig. 4 will accomplish the equivalent operation.

When the variable phasing capacitor in the circuit of fig. 4 is set at maximum capacitance, it becomes parallel resonant with its shunting inductors, thereby simulating an open circuit. At minimum setting of the phasing capacitor, the series capacitors are trimmed to produce a condition of series resonance, thus simulating an open quarter-wavelength of terminating line. In this manner you obtain a controlled phase shift of up to 180 electrical degrees by merely turning a knob.

Line lengths and coupling capacitances were first computed for a frequency of 7.2 MHz. Assuming a velocity of propagation factor of 0.66 for the RG-58/U cable chosen for the purpose, the transmission lines are each 135 inches (344cm) long. Coupling capacitors for 53.5-ohm coaxial cable are 414 pF. A pair of 220-pF silver mica capacitors were used in parallel. Alternatively, a 330 pF and a 100 pF capacitor were used for a second hybrid, with satisfactory results in each case. (If the system is built with RG-59/U coaxial cable for operation at the 70-ohm level, the capacitors will be about 316 pF.)

In order to determine the length of the transmission line a bit more accurately, a quarter-wavelength of line was fitted with a very small, single-turn coupling loop at one end. This assembly was then carefully grid dipped to frequency, the dipper frequency being checked on the receiver. Small bits of coaxial line were then clipped from the open end of the cable to raise the resonant frequency to the desired value. It is a great deal easier to cut a piece off than to fasten it back on, so be careful if you choose to go this route.

After the correct line length has been obtained, remove the coupling loop, double the line back on itself and cut it in two equal halves. Peel back as little covering and braid as necessary to make connections to the ends of the cables.

To check the hybrid, ports 3 and 4 are first left as open circuits and a

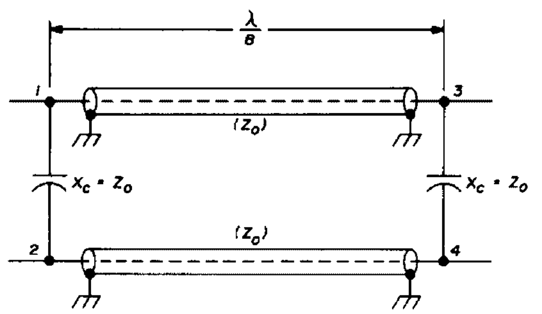


fig. 2. Capacitively-coupled hybrid is used in the phasor which controls the phased vertical antennas.

matched load is placed at port 2. An input vswr reading is now taken at various frequency intervals across the band to determine how well the line lengths have been chosen. At the center frequency a vswr of 1.05:1 or less

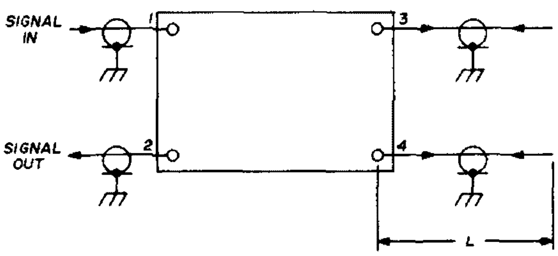


fig. 3. Quadrature hybrid terminated by equal lengths of transmission line.

should be possible if the line lengths and coupling capacitors are correct.

The isolation inherent in the hybrid can be checked by terminating ports 3 and 4 with matched loads, and the output at port 2 measured and compared with that obtained when the loads are *not present*. A *pair of* 51-*ohm carbon* composition resistors were used for this rough check, with the receiver at port 2 used as an indicator. Isolation of 20 to 30 dB was considered to be adequate for the purpose.

Once the hybrid has been checked out satisfactorily, the next step is the alignment of the reactance network. In the first place, the two inductors should be made as closely identical as possible. Both inductors should resonate at the center frequency with the phase control capacitor fully meshed. The grid dipper becomes necessary for this operation. Turns can be spread or squeezed together, and fastened with coil dope.

Once the coils have been trimmed, the phase control capacitor is turned to its *minimum* setting, and the series capacitors grounded, so that they are temporarily in parallel with the inductors. The series capacitors are now set so that the inductor-capacitor combination is again resonant at the same frequency of interest. Disconnect the temporary ground connection, and connect the series capacitors to ports 3 and 4 of the hybrid.

Connect a matched load to port 2 again and check the *input vswr* of the complete phasor. As the phase control capacitor is rotated through its range there should be no change in vswr.

In coupling to two separate antennas, a second hybrid, identical to that used in the phasor, appears to be the best arrangement as shown in **fig. 5**.

operation

To ensure placing a maximum signal lobe accurately in the direction of a given station, a pair of coaxial relays were hooked up as a reversing switch. The other station is first nulled out as far as possible, then the relays are actuated so that the connections to the transceiver and dummy load are reversed (fig. 5). This places a maximum signal lobe towards the selected station.

It can be seen that if a given station can be completely nulled out, the entire received signal is being routed to the dummy load, rather than to the receiver. Reversal of the order of connection by means of the relays then routes the entire signal from both *antennas to* the receiver. Since the effect is reciprocal, when transmitting both antennas combine to place a maximum lobe in the direction of the other station. It should not be difficult to calibrate the phase control in terms of direction so that this procedure can be omitted.

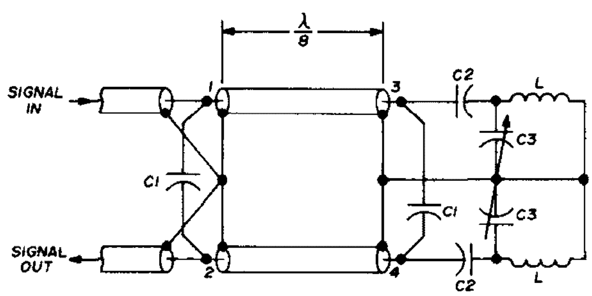
Don't expect a complete cancellation of the signal from a given station as no two antennas will have identical, perfectly circular radiation patterns. Maximum to minimum signal reports have varied from 1 to 6 S-units, with rare instances where the signal dropped completely into the noise level. If you consider that one antenna puts out a signal of 100 power units, while the other only puts out 95 units in a given direction, then the best possible cancellation

would still leave 5 units of signal. A ratio of 195 to 5 in power is about 14.1 dB, slightly more than 2 S-units. Therefore it should be apparent that the two antennas should be made as identical as possible.

Another factor in the operation of a system of this type is the spacing between the two antennas. If they are close together, there will be coupling present which will greatly affect the power distribution between them. With a spacing of around 120 feet (37m), the input vswr of the system varies over the phasor range from 1.0:1 to 1.2:1.

While any two relatively identical vertical antennas should perform satisfactorily, the pair I used were less than 15 feet (4.57m) in height, resting on a glass jar set on the ground. As this was a temporary experimental installation, a rather skimpy ground system of eight radials, each 30 feet (10m) long, was used. Matching was accomplished with an L-network.

It appears that the phasing system could be applied to two colinear horizontal dipoles. In addition to being, in many cases, easier to erect, the end-on arrangement of the radiating elements might be expected to show lower mutual coupling between the elements.



C1 hybrid coupling capacitors

C2 reactance network series capacitors

C3 phase-control capacitor (two 140-pF sections)

L approximately 14½ turns, 1¼" (32mm) diameter, 1¼" (32mm) long

fig. 4. Circuit for the 7-MHz phasor, showing grounding arrangement for the outer coaxial connections. Setup and test of this circuit is discussed in the test.

While not giving a full 360° coverage as uniformly as vertical antennas, the horizontal system might offer an interesting area for investigation.

It is felt that the results obtained with this experimental (and admittedly temporary) installation justify the con-

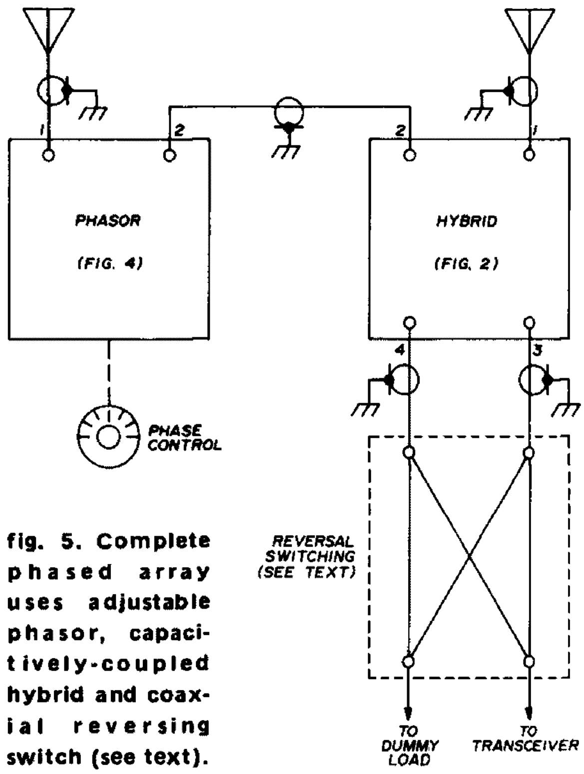


fig. 5. Complete phased array uses adjustable phasor, capacitively-coupled hybrid and coaxial reversing switch (see text).

cept of omnidirectional control of the antenna pattern. Although the wide spacing of the elements of the antenna system results in more lobes than a beam, the total area covered by the lobes should be about the same, with no loss of signal strength in directions of maximum signal as a result of the increased spacing. A vertical antenna system of this kind should offer much to the low-frequency DX man by combining the low-angle radiation of the vertical antenna with control of the directivity.

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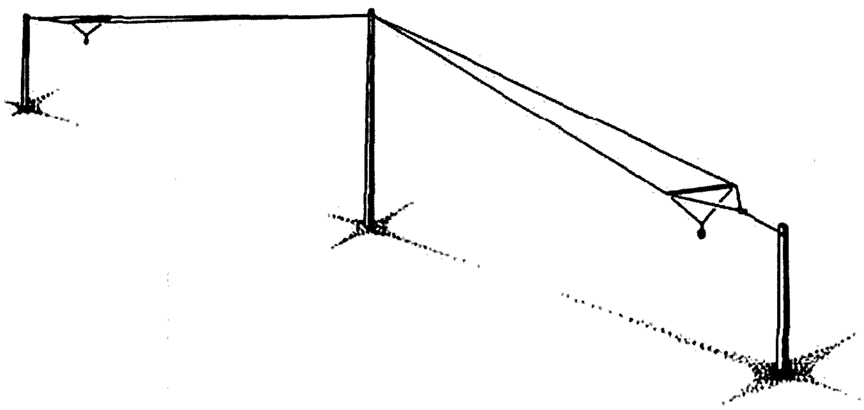


wide bandwidth bow-tie antenna for eighty meters

Discussion of a bow-tie antenna design using galvanized-steel wire that provides low swr performance over the entire 80-meter amateur band

Dwight Borton, W9VMQ, Box 93, Zanesville, Indiana

One of the problems requiring a decision when you establish a station is the selection of an antenna or group of antennas for the bands on which you want to operate. To make the best use of your environment and space available requires some study and planning.

In the case of the limited space of a city lot, especially for an 80-meter antenna, special steps may be needed for satisfactory results. A major consideration for an 80-meter antenna is the portion of the band you intend to use. The ordinary single-wire horizontal dipole, when used on this band, for example, will not work well over the entire band without some sort of antenna tuner or matching system. Verticals have the same limitations, but this article is concerned only with a horizontal antenna.

A typical horizontal dipole, resonant at 3.75 MHz, will approximate a series-

resonant circuit as shown in fig. 1. Resistance R represents the radiation resistance which will be about 50 ohms. The inductive and capacitive reactances, X_L and X_C, with a Q of 14 will be about 700 ohms (14x50 ohms) at the resonant frequency.

Fig. 2 shows a swr vs frequency curve for a horizontal, single-wire antenna, resonant at 3.75 MHz, measured and used as a standard of comparison for the experiments that follow. Tests made at four other amateur stations show the curve of fig. 2 to be typical for antennas of this type.

One important fact must be noted at this point: the majority of swr bridges made for amateur use will not provide accurate readings on the 80-meter band because of the non-linearity of the germanium diodes used in the simple bridge circuits. A Heath HM-102 swr bridge was used to obtain the curves presented here. This unit checked very closely with a Waters 365A reflectometer as well as with a standard Bird wattmeter. The typical, simple bridge will measure swr as much as 35% low on 80 meters.

Using the values of fig. 1, the inductive reactance of the antenna at 4 MHz will be (4/3.75)700 = 747 ohms; the capacitive reactance, (3.75/4)700 = 656 ohms. The net reactance is 747 - 656 = 91 ohms (inductive). At 3.5 MHz the inductive reactance will be (3.5/3.75)700 = 653 ohms; capacitive reactance is (3.75/3.5)700 = 750 ohms, and the net reactance is 750 - 653 = 97 ohms (capacitive). Neglecting the resistance change with frequency, which is small, the impedance at 4 MHz will be approximately 50 + j91 ohms; at 3.5 MHz the impedance will be approximately 50 - j 97 ohms.

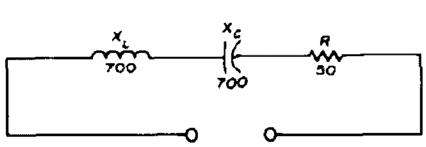


fig. 1. Equivalent circuit of single-wire horizontal dipole. Values are typical of those at resonance.

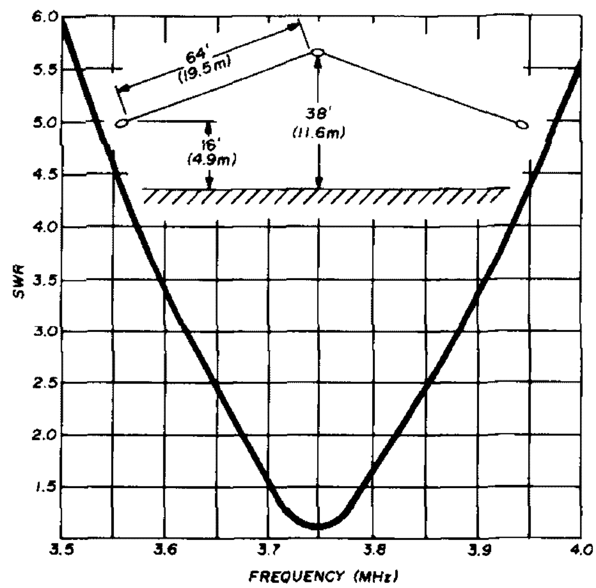


fig. 2. Swr vs frequency curve for single-wire antenna which is resonant at 3.75 MHz (radiation resistance = 45 ohms). Antenna is made from number-14 copper wire.

Fig. 2 shows the swr variations across the 80-meter band for the single copper-wire antenna with the maximum points at about 5.6:1 on the band edges. To reduce the swr (to broaden the frequency range of the antenna) you can reduce the Q by reducing the reactance or raising the radiation resistance. One method of reducing the reactance is by using a larger diameter antenna conductor. However, in most cases this is impractical at low frequencies. To reduce the Q of the single-wire antenna from 14 to 10, for example, would require a diameter of 640 mils or 0.64 inch (16mm). The use of RG-8/U coaxial cable, with the inner and outer sections

connected in parallel, reduced the Q of the antenna to about 12.

A much more practical method of increasing antenna bandwidth is by using the bow-tie or fan configuration shown in fig. 3. With a wire separation of 7 feet (2.1 meters) or more and copper wire, the Q can be reduced to about 10. This brings the swr at the band edges down to about 3.8:1. For swr no higher than 2:1, the bandwidth is increased to 190 kHz.

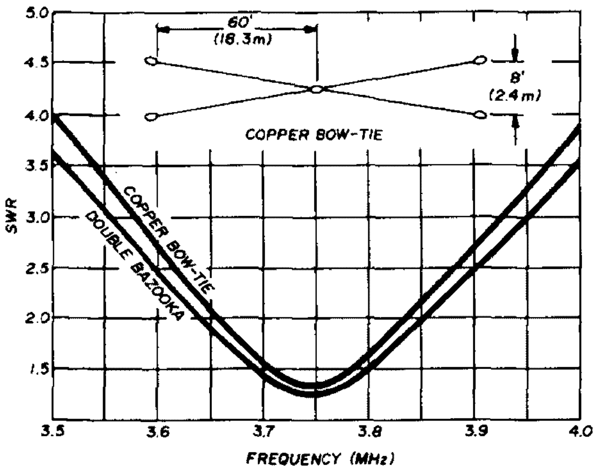


fig. 3. Swr comparison of the copper bow-tie and double bazooka antennas on 80 meters. Radiation resistance of the bow-tie is 35 ohms at 3.75 MHz; radiation resistance of the double bazooka is slightly lower.

double-bazooka antenna

The so-called double-bazooka or coaxial antenna is another modification for increasing the bandwidth of the basic horizontal dipole. However, the results were disappointing in the tests I made with this system. With new RG-58/U cable and very careful construction, with open-wire line for the end sections, the best I could obtain was an swr of about 3.5:1 at the band edges, an 8% improvement over the bow-tie (see fig. 3). The use of RG-8/U or RG-11/U for this antenna was not tried.

On the basis of the considerably greater cost and work required to build the double-bazooka antenna, it compares poorly with the bow-tie. The possible balun characteristic it is supposed to have is difficult to determine and of doubtful value.

galvanized wire

About two years ago, with more antenna experimenting in mind and copper wire in short supply, I obtained a roll of galvanized steel electric-fence wire at a farmers' supply store. When I built a single-wire dipole with this wire, a considerable lowering of swr was noted. The same wire in a bow-tie showed an swr of about 2.5:1 at the 3.5- and 4-MHz band edges. When I checked the radiation resistance at resonance, it was found to be about 50% higher than with copper, or about 75 ohms for a single wire and 50 ohm for the bow-tie (see fig. 4).

Speculation as to the reason for the increased radiation resistance, as well as how much antenna loss may have increased because of the higher resistance of this wire, led to quite a bit of research in reference books and experimenting.

First off, the wire I used is designated as number-16, but this refers to the

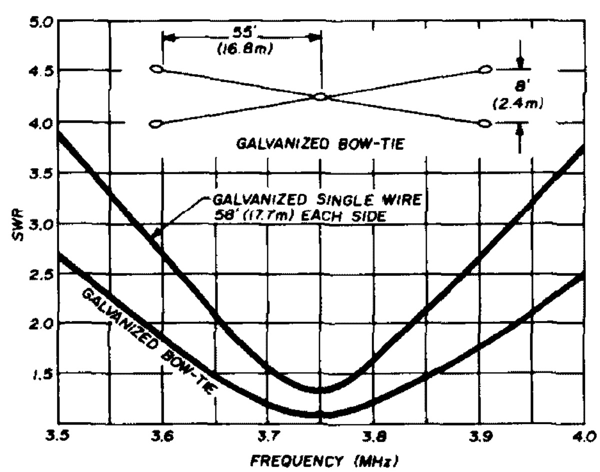


fig. 4. Swr performance of a galvanized single-wire dipole and a galvanized-wire bow-tie. Both antennas are resonant at 3.75 MHz. Radiation resistance of single-wire dipole is 75 ohms; radiation resistance of bow-tie is 50 ohms.

size before the zinc galvanizing is applied. Checking the wire table shows the diameter of number-16 as 50.8 mils (1.3mm). When checked with a micrometer, the wire measured 62 mils (1.5mm). The diameter of number-14 wire is 64 mils (1.6mm) so this wire is very nearly the equivalent. Although the use of galvanized wire for an antenna is by no means anything new, information on its rf characteristics is difficult to find. After failing to find anything in the antenna reference books, some experimenting led me to results that indicate, I think, a loss figure that is not too high when compared to copper.

The resistance tables show zinc with two to three times the dc resistance of copper. And, because of skin effect, most of the antenna rf current will be in the zinc coating. In an effort to get a comparison, equal lengths of number-14 copper wire and number-16 galvanized-steel wire were wound on identical forms and checked for Q at 3.75 MHz with a Q-meter. This test showed the copper-wire coil had a Q about six times that of the galvanized-wire coil. From this data it was assumed that a 6-to-1 ratio of rf resistance was fairly correct at 3.75 MHz.

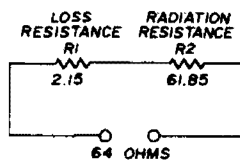


fig. 5. Relationship of loss resistance, R1, and radiation resistance, R2.

The actual loss resistance of a copper-wire antenna at 3.75 MHz is another thing that is very hard to find in the reference books. The loss resistance is usually considered to be "extremely low" or "negligible," and the only book I could find with anything like a definite statement was *Transmission Lines, Antennas and Waveguides*.[1] On pages 113 and 114 the authors stated that, when using 80-mil (2mm) copper wire at a frequency of 3 MHz, a dipole with 64 ohms load resistance at resonance will have 3% of the 64 ohms as loss resistance. This works out to be 1.92 ohms, and should be nearly the same at 3.75 MHz as the shorter length would just about balance the effect of the higher frequency. For number-14 copper wire, the smaller diameter should raise this to about 2.15 ohms $(1.92\sqrt{80/64})$.

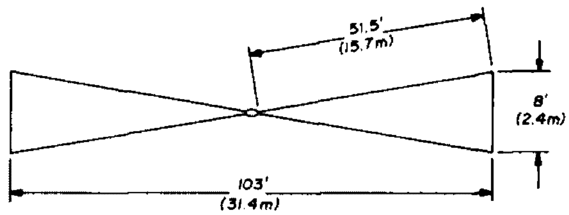


fig. 6. In this modification of the basic bow-tie antenna, the ends are tied together. Bandwidth is affected little by spreads from 6 to 16 feet (1.8 to 4.9 meters). Radiation resistance of this antenna is 50 ohms at 3.75 MHz.

Fig. 5 shows the relationship of the loss resistance and the radiation resistance. The percentage of power lost in heating the wire, of the total applied to the antenna, would be $R1/(R1 + R2) = 2.15/64 = 3.4\%$; the percentage radiated would be $100-3.4 = 96.6\%$. For the galvanized-steel wire, we can assume the loss resistance to be six times 2.15 or 12.9 ohms. It is assumed that this figure for the single galvanized wire will remain substantially the same with average variations in antenna height and inverted-vee angle. For the single galvanized wire, the loss ratio is $12.9/75 = 17.2\%$, yielding an efficiency of 82.8%.

Because the two wires of the bow-tie are in parallel for the antenna current, the effective loss resistance should be about one-half that of the single wire. For the copper and galvanized bow-ties, therefore, the loss resistance should be about 1.08 ohms and 6.45 ohms, respectively. The relative frequency response of the antennas that have been checked

can be expressed by the bandwidth over which they can be used with no more than a 2:1 swr (see table 1). The figure of 2:1 is used because this is the maximum swr specified by many manufacturers for their transmitters or transceivers. An swr of 2:1 is also the value above which line loss begins to mean something.

bow-tie antennas

A useful modification to the basic bow-tie antenna is that of tying the two wires together at the ends as shown in fig. 6. This shortens each side by about one-half of the end separation as shown. The overall length of 103 feet (31.4 meters) is very desirable where space is limited. The end connection can be made by using light-weight aluminum spreaders, a wood spreader with wire connector, or any other method that provides the mechanical spread and the electrical connection.

The spread of the wires can be either horizontal or vertical. A variation of spread from 6 to 16 feet (1.8 to 4.8 meters) makes very little difference in the swr characteristic. A very wide spreader, however, results in a proportionate reduction in overall length. Fig. 7 shows one arrangement that worked very well with swr performance slightly better than the standard spread.

table 1. Bandwidth of different 80-meter antennas for maximum swr or 2:1 (antenna resonant at 3.75 MHz).

antenna type	load resistance	loss resistance	percent loss	bandwidth
Single copper wire	45 ohms	2.15 ohms	4.8%	165 kHz
Single galvanized wire	75 ohms	12.90 ohms	17.2%	188 kHz
Copper-wire bow-tie	30 ohms	1.08 ohms	3.6%	190 kHz
Double bazooka	35 ohms	—	—	206 kHz
Galvanized-wire bow-tie	50 ohms	6.45 ohms	12.9%	325 kHz

A simple and economical L-network tuner, as in fig. 8, will allow an antenna cut for resonance at 3.75 MHz to be used over the entire 80-meter band with no more than 1.5:1 swr at the transmitter terminals. The maximum swr on the feedline is only about 2.6:1 so with RG-8/U feedline, the line loss due to swr would be about 0.34 dB at 3.5 MHz and 0.44 dB at 4 MHz, practically negligible amounts.

A comparison of the associated feedline loss of the two bow-tie antennas (no tuner) is presented in table 2. The larger feedline loss of the copper bow-tie is the result of the low radiation resistance of 30 ohms which results in considerably higher line current. This effect was checked out experimentally with a dummy antenna on the bench.

table 2. Losses of copper and galvanized bow-tie antenna systems with RG-8/U coaxial feedlines.

antenna type	frequency	antenna loss	feedline loss	total loss
Copper-wire bow-tie	3.75 MHz	0.17 dB	0.50 dB	0.67 dB
Galvanized-wire bow-tie	3.75 MHz	0.65 dB	0.28 dB	0.93 dB
Copper-wire bow-tie	4.00 MHz	0.17 dB	0.88 dB	1.05 dB
Galvanized-wire bow-tie	4.00 MHz	0.65 dB	0.40 dB	1.05 dB

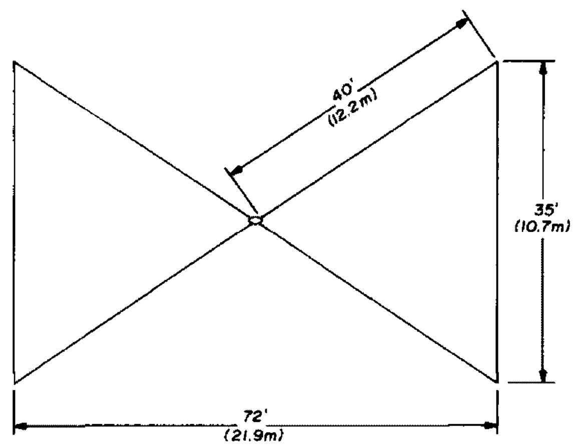
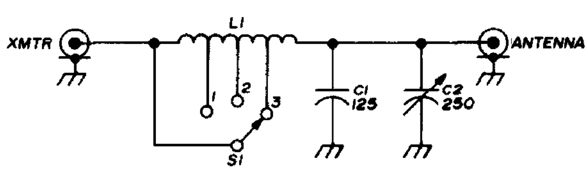


fig. 7. Bow-tie antennas can also be built with a very wide spread as shown here. As well as shortening the overall length of the antenna, this slightly improves the swr performance. Radiation resistance of this antenna is 50 ohms at 3.75 MHz.

Table 2 shows that the price paid in increased loss from the use of the galvanized steel wire is small enough to justify its use for the resultant broadband characteristic of the antenna. Its lower cost, compared to copper, is a fringe benefit. The durability of this wire, if my case is typical, is very good. The same wire has been up for two years with no visible rust. A coating of varnish or lacquer could be applied before putting the antenna up, if desired.



C1	125 pF mica, 1000 working volts
C2	250 pF air variable, 0.030" (0.8mm) spacing or greater
L1	9 turns no. 16, 1-7/8" (48mm) diameter, 1" (25mm) long, tapped at 2, 5 and 7 turns

fig. 8. Simple L-network antenna tuner which can be used to match the bow-tie antenna to 50 ohms over the entire 80-meter band (swr - 1.5:1 or less). Inductor L1 is 9 turns no. 16 airwound on 1-7/8" (48mm) diameter, 1-inch (25mm) long, tapped at approximately 2, 5 and 7 turns. Capacitor C2 should have spacing of 0.03" (0.8mm) or more.

operating Q

The operating Q of the bow-tie antenna can be reduced further by using the parallel compensating circuit shown in fig. 9. This circuit is simple, inexpensive and will work with any dipole to some extent. It makes use of the principle that a parallel tuned circuit has the opposite reactance variation on each side of resonance as that of a series

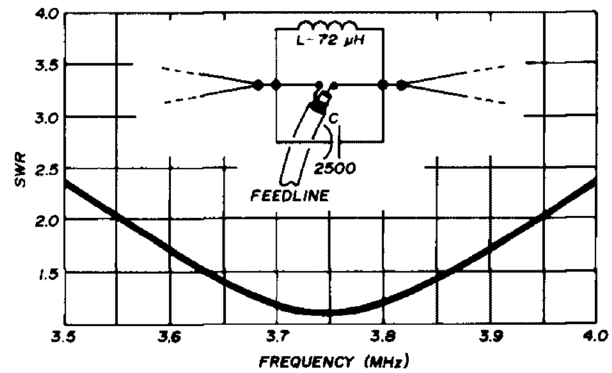


fig. 9. The bandwidth of the galvanized-wire bow-tie antenna can be increased still further by the addition of the parallel L-C circuit shown here (see text). Feedline is 88-feet (24.4m) long.

resonant circuit. When connected as shown in fig. 9, the parallel network will cancel some of the series reactance exhibited by the antenna on each side of resonance. The values of inductance and capacitance were determined by experiment for best results. The capacitance is 2500 pF and the coil is adjusted by means of a grid dipper to resonate with the capacitor at 3.75 MHz.

By shortening or lengthening this antenna, the resonant point can be moved higher or lower to obtain the desired coverage. Eight inches (20.3cm) of change, in each wire, will produce about 80 kHz frequency shift.

Considerable time and effort have been spent to answer two obvious questions: Why does the bow-tie antenna exhibit broader response than a single

wire, and why does the use of galvanized steel wire show the same effect? As for the first question, one reference book indicated that the bow-tie arrangement effectively increased the conductor size. That does not satisfy me. The most logical explanation seems to be that the two parallel wires reduce the inductance while at the same time increasing the capacitance between the halves of the antenna and the ground. The reduction in reactance is great enough to more than compensate for the reduction in radiation resistance, resulting in lower Q.

As to why galvanized steel wire increases the bandwidth of the antenna, it is thought that the 60% increase in feed-point resistance at resonance (in the bow-tie), as indicated by an antenna noise bridge, must be the reason for the lowered Q. Part of this increase is due to the higher loss resistance, of course, and this has been calculated to be about 6.45 ohms while the actual increase is 20 ohms. This would require a ratio of 20/1.08 or about 18.5 times as much rf resistance in the galvanized as the copper antenna. Even considering the limits of the Q meter for making coil comparisons, this is much too great an error to believe possible.

It has been noticed, however, that the galvanized wire, for a given resonant frequency, is about 4% shorter than the copper wire. This could be explained, as suggested by WBØBHG, by a "velocity factor" effect of the current flow slowing down on the higher resistance wire.

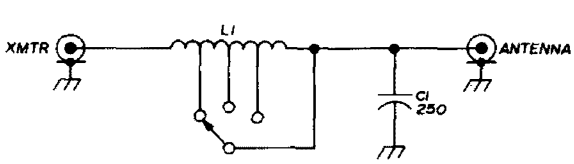
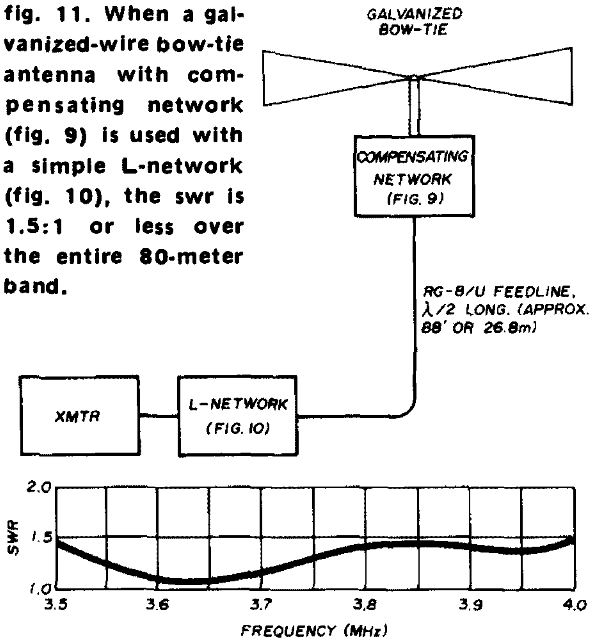


fig. 10. L-network for use with the bow-tie antenna with compensating network shown in fig. 9. Resultant swr curve is plotted in fig. 11. L1 is 9 turns no. 16, 1-7/8" (49mm) diameter, 1" (25mm) long, tapped at 2, 5 and 7 turns. Capacitor C1 is a 250 pF mica, 1000 working volts.

fig. 11. When a galvanized-wire bow-tie antenna with compensating network (fig. 9) is used with a simple L-network (fig. 10), the swr is 1.5:1 or less over the entire 80-meter band.



This might raise the radiation resistance and lower the Q.

One experiment was tried using number-26 copper wire instead of galvanized steel. The swr at the band edges of 80 meters was about 3.2:1 and the radiation resistance about 65 ohms. The rf resistance of number-26 copper, according to the wire table, should be about 85% of the single galvanized-steel wire.

The transmission line used for all tests and swr curves was 88 feet (26.8 meters) long, checked out with a noise bridge for one-half wavelength at 3.75 MHz. It was found experimentally that the circuit of fig. 10, with the values shown, when inserted between the transmitter and feedline, modified the swr curve to that shown in fig. 11. For this result, however, the transmission line must be close to one-half wavelength long. To obtain the averaging out effect the transmission line should be within 5% of one-half wavelength long.

The simple circuit of fig. 10 replaces that of fig. 8 when the parallel compensating circuit at the antenna and a half-wave feedline are both used. This brings the system to the point where no tuning

is needed at all, with very low swr at the transmitter output terminals.

operation

The ability to work across the entire band with no more than 1.5:1 swr, as shown in fig. 11, provides very smooth operation. A variable antenna tuner with a coax fed copper antenna will also cover the 80-meter band but with a much more complicated tuner and with much higher line loss.

Swr curves were run with and without a balun at the center of the antenna. Both straight-core and toroid types were tried. The only observed difference was a downward shift in resonant frequency by about 50 to 75 kHz. Substitution of number-14 galvanized wire with an actual diameter of 75 mils (1.8mm) resulted in a very small change as compared to number-16 wire. A third wire, strung between the two wires of the basic bow-tie, was tried with very little change.

One thing I did notice was that twisting the two wires together at the center, even for 2 or 3 feet (61 to 91cm), raised the swr about 6%. This effect led me to try bracing the two wires about 10 inches (25.4cm) apart at a point about 18 inches (45.7cm) out from the meeting point, but no improvement was observed.

I experienced no difficulties from the wires getting twisted or tangled after the antenna was installed. Winds up to 60 mph (97 kmh) have given no trouble.

All experimental work was carried on jointly with W8URR who first suggested the use of the bow-tie arrangement with which he had already done considerable experimenting. Both he and W8SAY are using the antenna with very satisfactory results.

reference

1. King, Mimnow and Wing, *Transmission Lines, Antennas and Waveguides*, McGraw-Hill, New York, 1945.

ham radio

loop antenna

receiving aid

A method for improving signal-to-noise ratio on the lower-frequency bands

Ken Cornell, W2IMB, P.O. Box 721, Westfield, New Jersey 07091

You've heard it before and it's worth repeating: "If you can't hear 'em, you can't work 'em." Most hams I've talked to or heard on the air on the lower frequencies use quarter- or half-wave antennas in vertical, horizontal or inverted-vee configurations. While these antennas do a creditable job of transmitting rf, they can be noise collectors for receiving because of their physical size and proximity to man-made noise sources.

Presented here is a scheme that uses a fet preamp or a Q multiplier with a loop antenna which is easy to rotate from the operating position and will allow you to discriminate between signals and noise. The loop is only 18 inches (46cm) on a side with two windings spaced 2 inches (5cm) apart. A complete parts list and construction drawings are included. The loop structure can be made with a few pieces of wood, and the electronics are simple and inexpensive.

loop antennas

The rotary beam, if designed properly, discriminates well between the desired signal and interference. However, not many amateurs are fortunate enough to have a rotary beam for the lower amateur frequencies. If we sacrifice some gain and use a small loop that will give reasonable attenuation on an interferring signal or noise, we'll have a desirable receiving aid. Such an antenna has been around for years and its characteristics are well documented.[1,2]

Largely ignored by today's amateurs, the loop antenna was a standard piece of equipment in early ham stations but was generally huge and cumbersome because of the low frequencies in use in those days. The loop doesn't have to be large to be an effective noise discriminator on the lower amateur frequencies in use today. The entire assembly, including antenna and preamp or Q multiplier, can be mounted on your receiver. Used properly, this combination could mean the difference between solid reception and frustration.

Loops have a pattern that exhibits maximum response in the plane of the loop and minimum response in a plane normal to the loop (fig. 1). Note that

this response is exactly opposite to that of a quad antenna. This suggests that a small loop antenna with appropriate electronics might be useful as a noise discriminator on the higher amateur frequencies when used as a receiving aid with a quad.

receiver input impedance

A loop antenna with its tuning capacitor is a common LC circuit. It could be substituted for the tuned circuit in your receiver rf amplifier. Few hams, however, would want to make the circuit modifications. Most *modern receivers* use a low-impedance antenna input; and since the basic loop has a high impedance, we need a one-turn loop coupled to the large tuned loop to act as a low-impedance transformer to the receiver input. See **fig. 2**.

design examples

The first thought that might enter your mind when constructing a loop is to use the number of turns required for the lowest frequency desired, and provide taps in the winding for the higher frequencies, which is not an uncommon practice with coils. My experience using this means has proved that it is *not* practical. For some reason, the unused turns seem to saturate the loop, and a sharp resonance becomes hard to achieve. Also the directivity pattern becomes questionable. I've found that it's better to wind two separate loops on

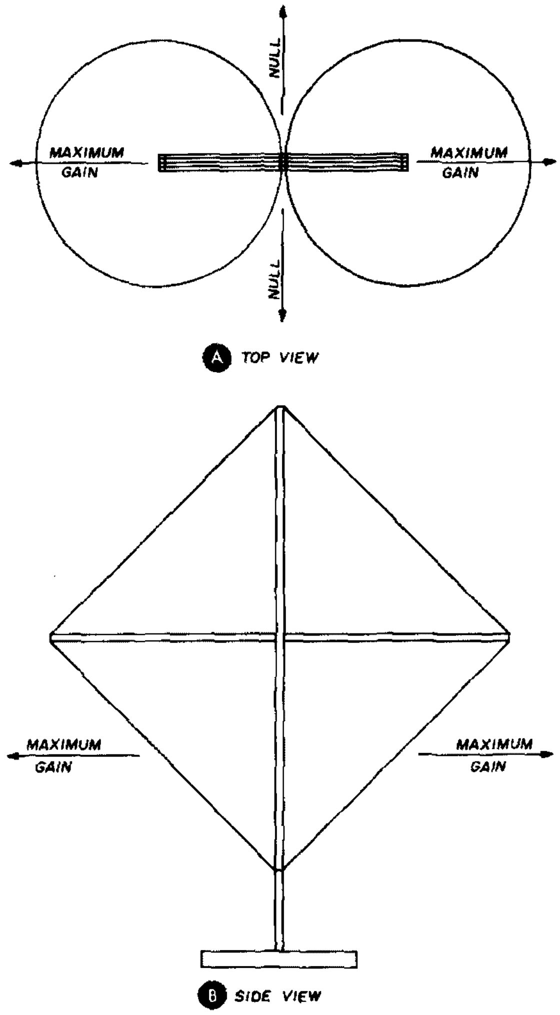


fig. 1. Loop antenna field strength patterns.

the same frame, with the most spacing you can provide between the two windings, e.g., about 2 inches (5cm) will provide excellent performance. I've also found that as far as the LC ratio is

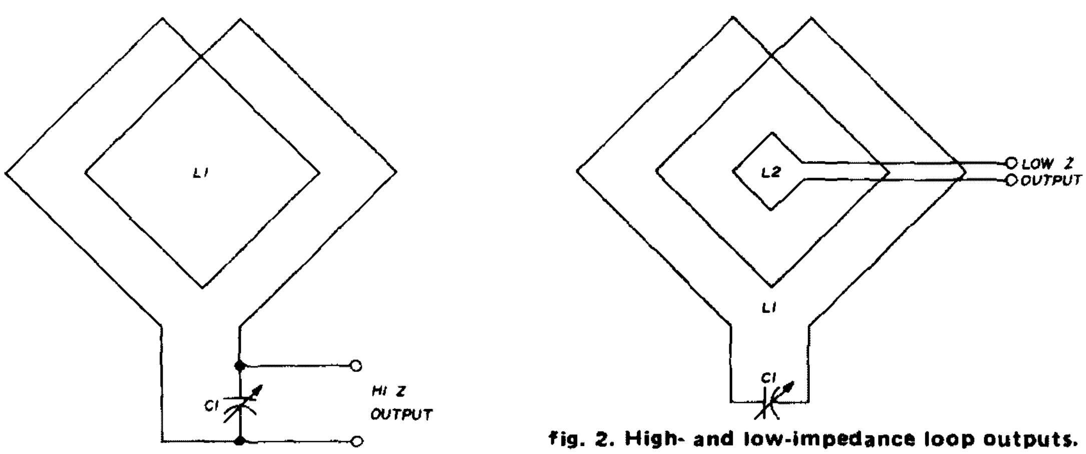


fig. 2. High- and low-impedance loop outputs.

concerned, the higher the capacitance (within reason), the more effective the loop performance.

An example of this is the loop that I will describe. It is approximately 18 inches (46cm) on a side and contains two windings spaced 2 inches (5cm) apart. One winding consists of two turns spaced 1/4 inch (6.5mm) and covers 40 and 80 meters. The other winding consists of 5 turns spaced 1/4 inch (6.5mm) and covers 80, 160 and the high end of the broadcast band. With this arrangement, the loop is more effective on 80 meters using the 2-turn loop with high capacitance than the 5-turn loop with low capacitance.

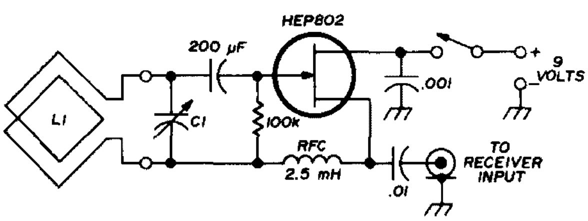


fig. 3. Fet preamp for the loop antenna.

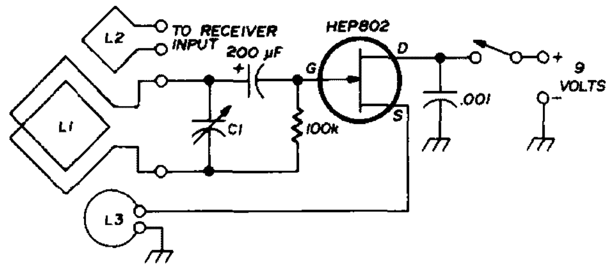


fig. 4. Loop antenna with Q-multiplier. Regeneration control is by means of L1, L3 coupling.

Loop with preamp input. While the loops performed satisfactorily in their basic configuration, as my first experiment I decided to add an rf amplifier using a solid-state device, and the circuit of fig. 3 evolved. Here the loop is connected to the gate of an HEP-802 fet and output to the receiver is taken from the source. Results were good, but I felt that the null could be improved, perhaps by increasing circuit Q.

Loop with Q-multiplier input. Recalling the principles of the Q multiplier, I decided to substitute such a circuit for

table 1. Parts list for loop antenna.

item	quantity	description	dimensions	use
1	4	wood strip	1/2 x 3/4 inch	loop frame
2	3	support	fig. 6A	loop wire
3	1	support	fig. 6B	bottom loop wire
4	2	loop frame brace	fig. 5	feedback loop L3
5	1	base	to suit	loop support
6	2	support bracket		loop frame
7	2	angle bracket		
8	2	circuit board	3 inches (76mm)	fig. 3 and 4
9	1	wood dowel	see text	feedback loop L3 support
10	1	wood strip	see text	support for ends of L3

The variable capacitor I use came from an old BC set. I wired its two stators in parallel and assume it has a maximum capacitance of 600 pF. I use a small dpdt switch to select the desired loop.

the fet preamp (see fig. 4). Feedback control is obtained by an adjustable loop of wire, L3, coupled to the loop, L1. Receiver input is taken from L2. By rotating L3 within the field of L1 the desired amount of regeneration can be

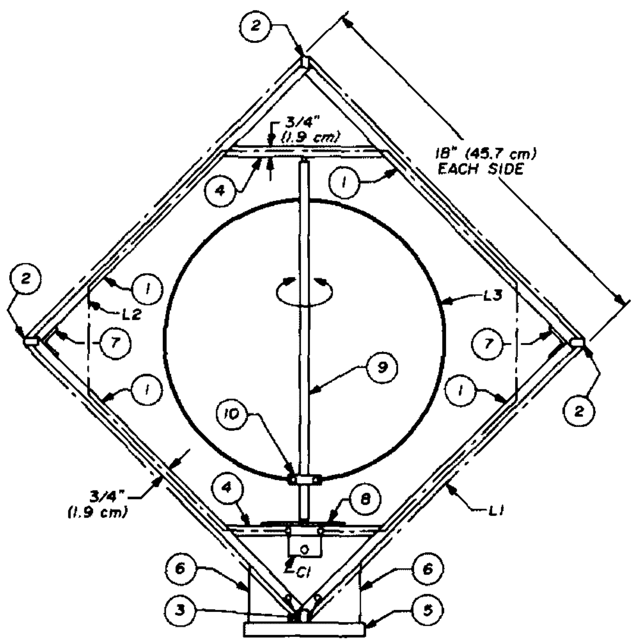


fig. 5. Construction details. Circled numbers refer to table 1. L3 and pieces 9 and 10 are required only if a Q multiplier is used.

obtained, circuit Q is improved, and the *null* is much more pronounced. Better control is obtained for separating noise from the desired signal.

Note that a loop antenna can also be used as an absorption wavemeter. When using it with solid-state devices beware of placing the loop in a strong rf field.

construction

Fig. 5 and the parts list (table 1) should be self-explanatory. The basic loop contains two windings: a two-turn loop for 40 and 80 meters and a 5-turn loop for 80, 160 and the high end of the broadcast band. Four turns can be used if operation on the broadcast band is not desired.

The structural support for the loop wire is made from two strips of wood fashioned in a cross. The simplest winding is made by starting the wire from the outside, winding around and around, to the inside. The wire is supported on the cross arms with tacks or notches. The size of the frame, number of turns, and spacing will be determined by the desired frequency coverage.

Another winding pattern is made as a large, square-shaped, wide-spaced solenoid. In this case, short strips of wood are required at the appendage of the supporting arms, with notches to support the wire. Here again the size, number of turns, and spacing will depend on the frequency range desired.

Incidentally, there is a school of thought that reasons as follows: Since the maximum gain of a loop antenna is in the plane of the wires, a loop wound in a wide-spaced solenoid configuration and mounted in a diamond position (points of the square top and bottom) will have a better capture area. While I can't prove it, I am inclined to agree.

Since the wire is the only important element of a loop, the basic construction of the frame can be left to your ingenuity and materials on hand. The frame, of course, should be made with *insulating material*. I suggest that the wood frame be painted with coil dope.

Fig. 6 details the loop wire support arms. *Note that one extra notch for* wire is required in the bottom arm due to the winding pattern. I formed the notches in the support arms in the following manner. I drilled the required number of holes in the arms on a line

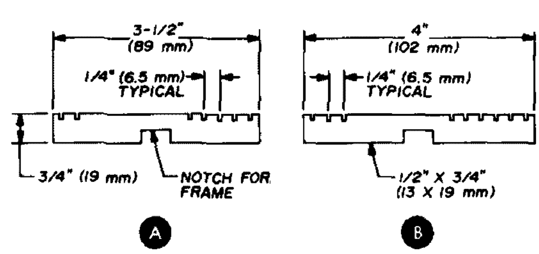


fig. 6. Loop wire support arms.

about 1/8 inch (3mm) in from the *outside edge*, then clamped the supports in a vise and cut slots into the holes with a hacksaw. Next I folded a piece of sandpaper and widened the slots to suit the wire size. A piece of string a little

larger than the wire diameter, rubbed over a candle and worked into the slots, made it easy to obtain proper tension in the loops when securing the ends.

The feedback-loop support dowel should be cut 1/4 inch (6.5mm) shorter

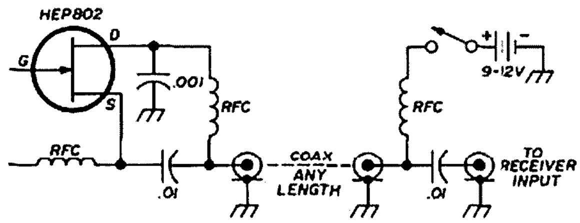


fig. 7. Suggested power feed for a remote loop with fet preamp.

than the space between the supports. The ends of the dowels should be drilled on the vertical centerline to take a short length of heavy wire. This wire should project 3/8 inch (9.5mm) at the top and 1/8 inch (3mm) at the bottom. The supports should be drilled at their centerline to accept the wire. The dowel is inserted into the top support first, then swung in to the bottom support hole. Rotation to obtain proper feedback is by manual adjustment. In constructing the loop frame, I cut all pieces to proper length and shape, then assembled them using *Pliobond* cement and a few brads.

Fig. 7 is a suggested circuit to supply power to a loop antenna preamplifier through the coaxial transmission line when the loop is in a remote location. I suggest that the voltage drop that might occur in the two rf chokes be checked and, if warranted, the voltage increased.

references

1. Frederick E. Terman, *Radio Engineers' Handbook*, McGraw-Hill, New York, 1943, page 813.
2. Keith Henney, *Radio Engineering Handbook*, McGraw-Hill, New York, 1959, pages 3-26, 19-7, 19-21, 19-116, 19-184.

ham radio

tilt-over tower

uses extension ladder

Tilt-over tower based on the use of a 40-foot aluminum extension ladder is inexpensive and easy to erect

Henry S. Keen, W5TRS, Fox, Arkansas 72051

It is difficult to beat the tilt-over tower as a means of supporting your beam antenna as it permits easy access to the elements for adjustments or repairs. Unfortunately, these structures are expensive and require a rather elaborate installation. If you are required to move because of a job change, retirement, or just plain restlessness, there then arises the problem of what to do with the tower. Many times it must be either given away, or sold at a substantial loss.

The system described here makes use of a readily obtainable aluminum extension ladder which is light, easy to handle, and far less expensive than the usual tower. And if you should move, it is easier to transport or, if necessary, dispose of the ladder which can always be put back to the sort of work for which it was originally intended.

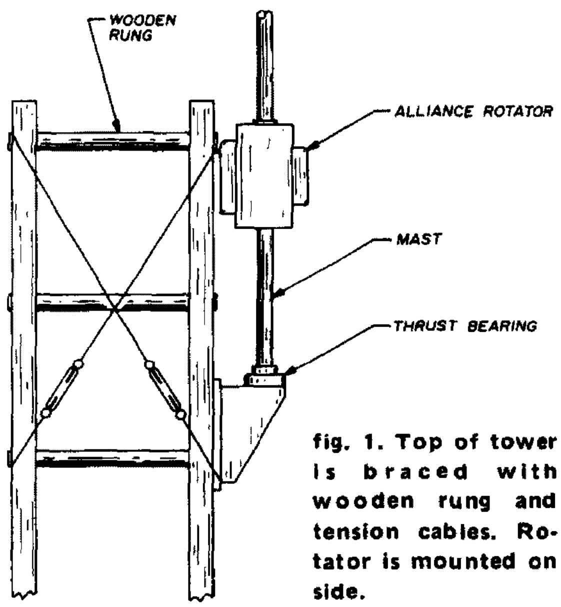


fig. 1. Top of tower is braced with wooden rung and tension cables. Rotator is mounted on side.

construction

After installation of this tower, the whole assembly is erected in the retracted position; the ladder is then extended to its full height for rigging. For my 15- and 10-meter quad an Alliance TV rotator with thrust bearing was mounted on the ladder as shown in fig. 1. To prevent twisting and damage to the ladder from wind action, a wooden rung was inserted at the extreme top of the upper section, with two crossed cables with turnbuckles for tensioning, as

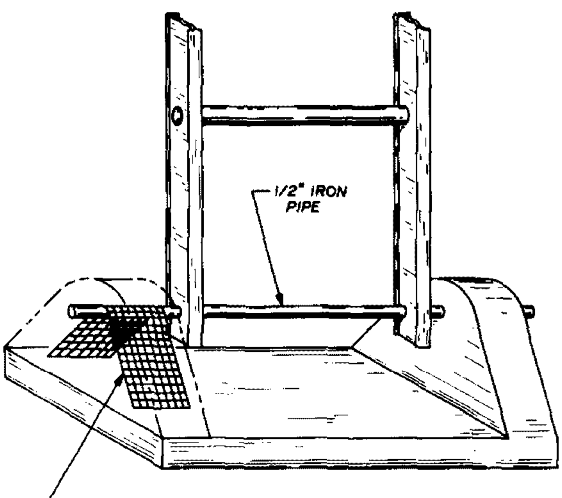


fig. 2. Hinged tilt-over base is made from *reinforced concrete.*

shown in fig. 1. Don't mount your beam without this modification to the ladder.

The swiveled feet originally installed on the bottom of the ladder were removed by drilling out the rivets and reaming the holes to a diameter permitting the insertion of a length of half-inch (13mm) iron pipe (fig. 2). This pipe becomes the pivot which anchors the bottom of the tower.

The ladder is secured vertically against the end of the house by a lag bolt through a board fitted between the vertical members of the ladder's lower section as shown in fig. 3. An eyebolt to the frame of the house secures a block

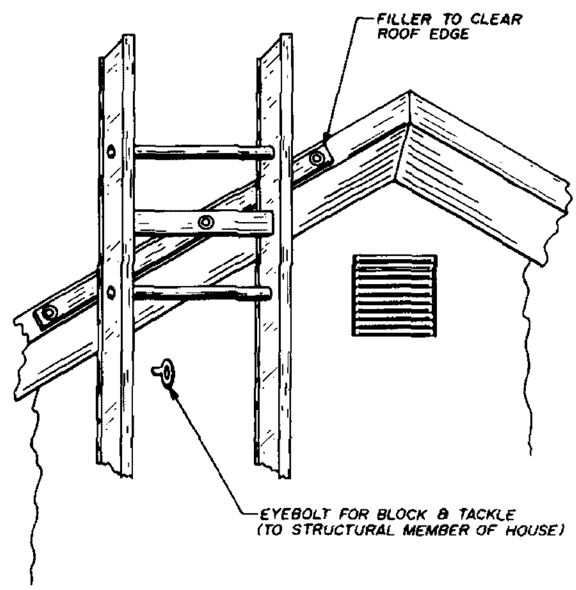


fig. 3. Method of securing the lower ladder section to the house.

and tackle which is used to raise and lower the assembly.

Guy wires are fastened to the ladder by running them through the hollow rungs, and attaching the end as shown in fig. 4. Thus, extreme or continued stresses will not tend to pull the ladder apart.

Ladders of this type are available in lengths up to 40 feet (12 meters). With the additional height provided by the 10-foot (3m) rotating topmast, which supports the quad, a total height of more than 45 feet (13.7m) can be obtained.

With this type of antenna support, as well as with more conventional towers, the problem of neighborhood children,

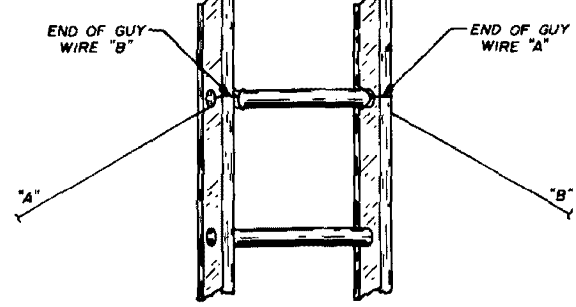


fig. 4. Fastening guy wires to the tower (details of fig. 1 omitted for clarity).

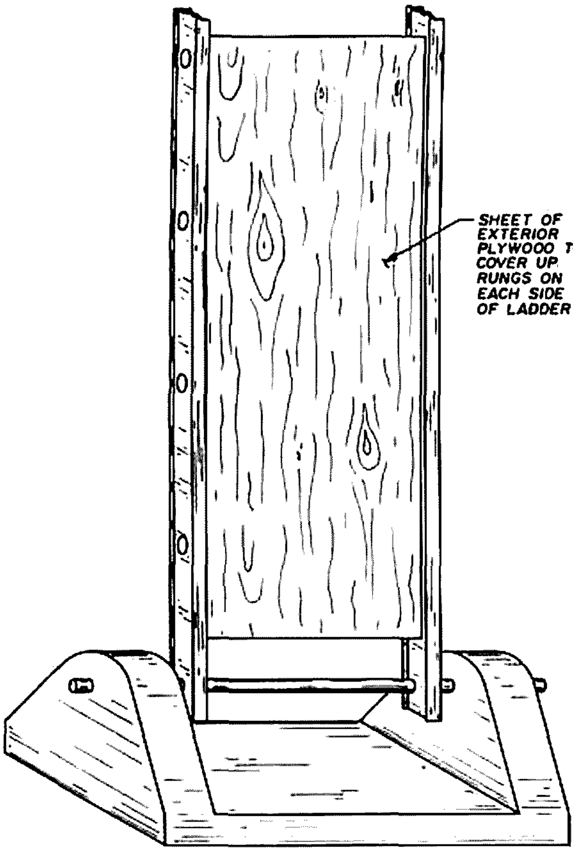
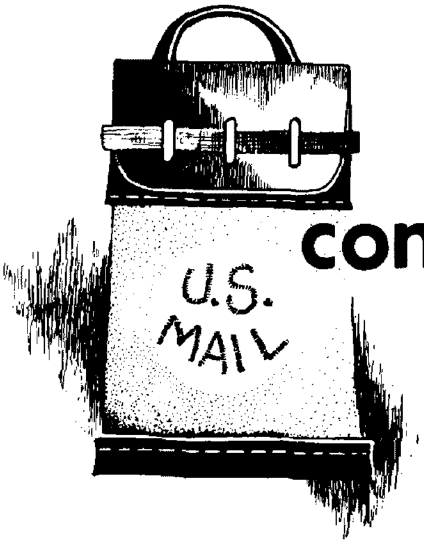


fig. 5. Extension-ladder tower can be child-proofed by adding plywood covers to each side of the ladder.

or your own for that matter, climbing it, with resulting danger of injury, is ever present. The ladder can easily be made "child proof" by using two pieces of outdoor plywood, one on each side of the lower section, so that the rungs are effectively covered up (**fig. 5**). A conventional tower can be protected in a like manner by proper location of plywood shields so that no hand-holds are present. Means of securing the shield will be left to the ingenuity of the builder. A hasp from one shield through the other, with a padlock, would seem to be an effective arrangement.

This installation has been in place for a year or so, and has been very easy to operate single-handedly, taking a little over an hour to bring down the antenna, make whatever adjustments are necessary and put it back into operation.

ham radio



comments

improved log-periodic beams

Dear HR:

When I originally started designing and building fixed log-periodic wire beams, I cut the active elements to resonate near the center of the 20- or 40-meter bands. I later discovered that it was better to cut these elements for the low end of the desired band. In my article on feed systems for log-periodics in the October, 1974, issue of *ham radio*, the rear element (element 1) was shown to be 35 or 36 feet (10.7m to 11m) long for the 20-meter antennas, and 70 to 72 feet (21.3m to 21.9m) long for the 40-meter versions. In these log-periodics the active element (element 2) was 32 to 33 feet (9.8m to 10.1m) and 64 to 66 feet (19.5m to 20.1m), respectively, for the 20- and 40-meter antennas.

I have since found that the forward gain and front-to-back ratio at the low-frequency end is improved by making elements 1 and 2, respectively, 37 feet (11.3m) and 33.4 feet (10.2m) for the 20-meter antennas, and 74 and 66.8 feet (22.6m and 20.4m) long for 40 meters. The remaining elements will be proportionately longer.

When designing log periodics from scratch (see article on page 14), I now use 13.9 MHz and 6.9 MHz, respectively, for the low cutoff frequency of 20- and 40-meter log-periodics. This results in improved performance at the low end while the midband and high-end performance remain the same.

When installing a log-periodic beam, be sure that the area in front of the beam is clear of obstructions or other antennas. If you position your standard dipole or test antenna too close to the front of your log-periodic, it may completely upset the pattern. This was called to my attention by Tony Mony, K4LD. He originally had a 7-MHz inverted-vee only 10 feet (3m) in front of his 7-MHz log-periodic (which showed no gain). When he removed the inverted vee, the log-periodic provided lower swr over the 7-MHz band as well as good forward gain.

George E. Smith, W4AEO
Camden, South Carolina

remotely-controlled gamma match capacitor

Have you ever wished you could adjust that gamma-matching capacitor on your beam from the ground? Or risked your neck up on a tower to get that last couple of tenths of vswr?

Here's a way to make that adjustment right in your shack while sitting down. Burstein-Applebee* has a surplus TV remote-control reversible motor for $3.95 (catalog number 18A1465) which has enough torque to do the job. The unit comes with a 10k pot with protrud-

*Burstein-Applebee, 3199 Mercier Street, St. Louis, Missouri 64111.

ing shaft for hand adjustment. I coupled this shaft to an E. F. Johnson 140-pF transmitting variable with an insulated flexible coupler, weatherproofed the whole thing in an aluminum box, and mounted it on the boom near the driven element. Since the Johnson capacitor has a double-ended shaft, a control knob can be added for hand adjustment if you insist. You need a small three-wire cable down to the shack for remote operation and Burstein-Applebee includes a schematic for the motor which operates on 115 volts, 60 Hz.

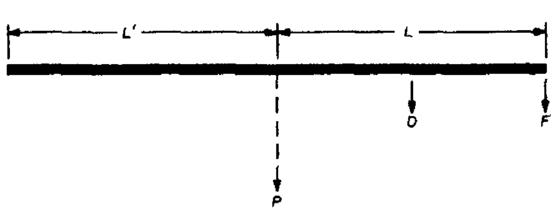
Since it's an instant start and stop device, with a little practice you can catch it at the bottom of the null. The 10k pot can also be used to work up a servo-control to operate a multi-position wafer switch for remote antenna switching. Disable the detent mechanism of the switch, though, as the motor doesn't have enough torque to overcome it.

Forrest Gehrke, K2BT

antenna mast design

Dear HR:

The technique of antenna mast design discussed in my article "Design Data for Pipe Masts," in the September, 1974, issue of *ham radio* is based on approximations accounting for the fact that pipe masts are highly flexible structures.



$$L'P = (L' + \frac{L}{2})D + (L' + L)F$$

SOLVE FOR P

WB9DES recommends that such masts be designed by the technique of analysis used for rigid beams. This can be done using the curves of the article as follows:

1. Calculate the length of the top section as in the example.

2. Calculate the second section length in the same way, but treat this as a trial value. For a second trial, refer all upper loads to the top of the second section: From the curves determine the allowable section length, and use this as a second trial value. Repeat until the trial value and graph value agree.

3. Repeat the process of step 2 for each lower section until the desired height is obtained.

Masts designed by this method will deflect less under load, and will have a larger safety factor.

R.P. Haviland, W3MR
Daytona Beach, Florida

speech processing

Dear HR:

My attention has been called to G6XN's article of speech processing in the November and December, 1972, issues of *ham radio* and the subsequent correspondence. I feel that there are still some points to be clarified, particularly in reference to *audio* speech clipping.

It is rightly supposed that *audio* clipping produces harmonic distortion products within the upper registers of the speech spectrum. The superiority of *rf* clipping comes from the fact that such products are largely eliminated. The unwanted distortion products can be regarded as noise and their existence constitutes a decrease in the signal-to-noise ratio, making it more difficult for the listener to perceive the essential speech components. Harmonic products are not, however, the only disturbances produced by *audio* clipping. Largely overlooked is a high-frequency mutilation effect. What happens is that on any extreme swings of the primary low-frequency components, any high-frequency ones get erased by the limiting action of the clipping diodes. This can be seen if one analyzes the result of the clipping of a compound waveform

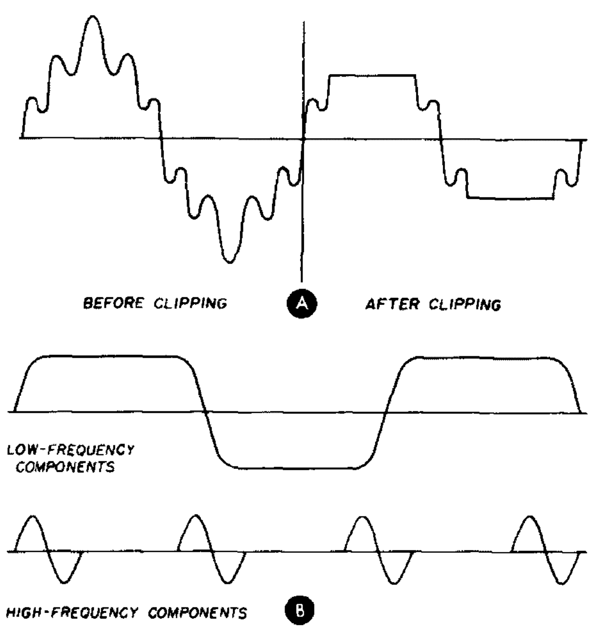


fig. 1. Speech clipping results in mutilation of the higher audio frequencies by the lower ones as shown here. VK4LR suggests a sub-band speech processing system to solve this problem (see text).

consisting of two components of about equal amplitude (see fig. 1A). Separating the two components after clipping (fig. 1B) illustrates that the high-frequency one has been reduced to modulated pulses. In other words, in the presence of a strong low-frequency component and with heavy clipping, any high-frequency component information can only be conveyed during and near the times of zero low-frequency component value. All high-frequency components thus become strongly mutilated. As the human voice relies for the most part on a number of simultaneously occuring frequency components, articulation is destroyed by this high-frequency mutilation as well as from the occurence of harmonic products caused by wave squaring. While *rf* clipping eliminates the harmonic product problem it does little to cure the mutilation of higher frequency audio components.

One solution I am investigating is that of splitting the speech band into a number of sub-bands and treating these separately by clipping and filtering. The sub-bands are combined after processing to reconstitute the full speech spectrum. Thus, any harmonic products generated by the lowest sub-band, 350 to 700 Hz, are filtered off by a lowpass filter section set to cut off above 70 Hz. Since the first troublesome harmonic of the lowest possible frequency that can be passed is at 1050 Hz, it is well into the filter rejection region and can cause little offense. Providing several components of the speech waveform do not both fall into the same sub-band, no high-frequency mutilation can occur. Analysis of typical speech pattern shows that the formats generally fall roughly an octave or so apart. Thus simultaneous occurrences of two formats in the same sub-band would be infrequent.

Preliminary test of the multi-sub-band system indicates that very little change in speech quality results when the system is introduced to the circuit. No substantial data has been yet obtained as to the communication gain that might be had, but some reports indicate that the gain is considerable.

L.R. Newsome, VK4LR
Queensland, Australia

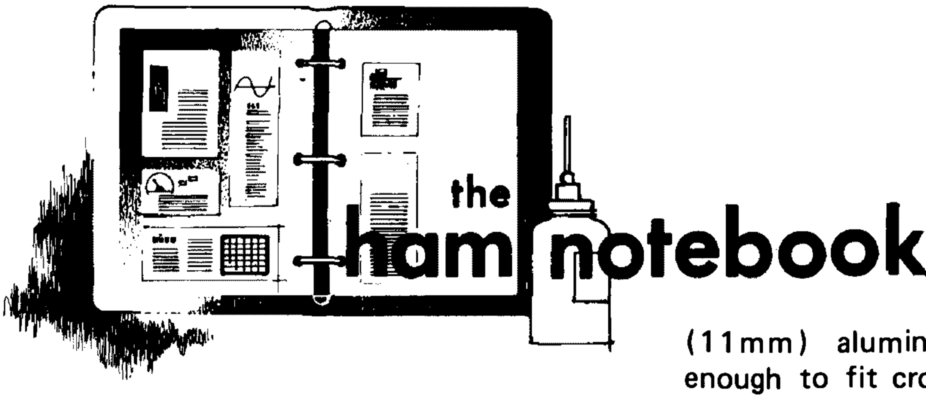
reciprocating detector

Dear HR:

I built a reciprocating detector described by W1SNN* and found that the concept of locating the reference filter in the middle of the passband was not acceptable, at least for receivers with high selectivity. To use it, I incorporated another stage of conversion in my Collins 75A3, using the original bfo for beating so that the passband could be shifted in relation to a new reference filter crystal frequency of 80 kHz. With this addition and a circuit modification to make the feedback symmetrical, performance of the circuit has been very worthwhile.

D.H. Gieskieng, W6NLB
Rialto, California

*Stirling Olberg, W1SNN, "Reciprocating Detector," *ham radio*, March, 1972, page 32.



unique method of clamping aluminum tubing

Have you ever tightened the U-bolts around a 1¼-inch (32mm) aluminum tube, only to have it slowly collapse as you torque up the nuts? To add to the frustration, the aluminum tube takes a permanent set which limits the amount of force the clamp can be made to apply.

The use of steel U-bolts to join tubing is common practice with commercial antenna installations. The U-bolts are simple and economical, but their use with aluminum tubing generally leads to the problem mentioned above. An alternative is to fabricate special clamps, such as those illustrated in the *ARRL Handbook*.[1] If minimizing weight on the mast is as important to you as it is to me, then the following description of a relatively simple fix, using commercial U-bolts, should be of interest.

Materials. The parts that are needed are easily made with a bench vise and a hacksaw. If a tube cutter is available, it can also be used to advantage. Parts to be fabricated are a half-section of 1¼-inch (32mm) steel tube about three inches (76mm) long, and two 7/16-inch (11mm) aluminum tubes just long enough to fit crosswise inside the 1¼-inch (32mm) tubing (about 1-1/8 inch [28.5mm] long). The latter should be a number-16 tube which has an inside diameter of 0.307 inch (8mm).[2]

Step-by-step procedure. First, using a hacksaw or tubing cutter, cut off a 3-inch (76mm) section of 1¼-inch (32mm) thinwall steel tube and drill a set of holes through it for the U-bolt.

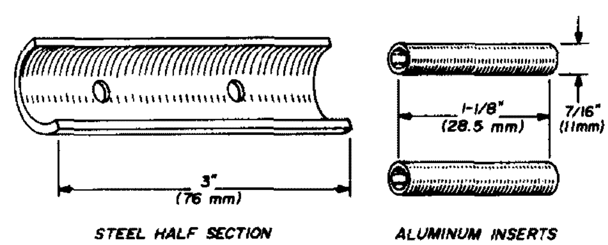


fig. 1. Mechanical parts required for clamping aluminum tubing without deforming it consist of two aluminum inserts and a steel half-section. Assembly of the parts is shown in fig. 2.

(The open end of ¼-inch [6.5mm] steel clamps are generally 1-3/4 inch [44.5 mm] apart.) Second, place the tube crosswise in a vise. Using a hacksaw, cut it in half lengthwise, midway between the two sets of clamp holes. Remove the burrs with a file.

The third step is to cut two lengths of 7/16-inch (11-mm) aluminum tubing

1. *The Radio Amateurs Handbook*, 50th Edition, ARRL, Newington, Connecticut, 1973, figure 22-18, page 634.

2. Lewis McCoy, W1ICP, "Aluminum Tubing —What Sizes Are Available?" *QST*, June, 1969, page 16.

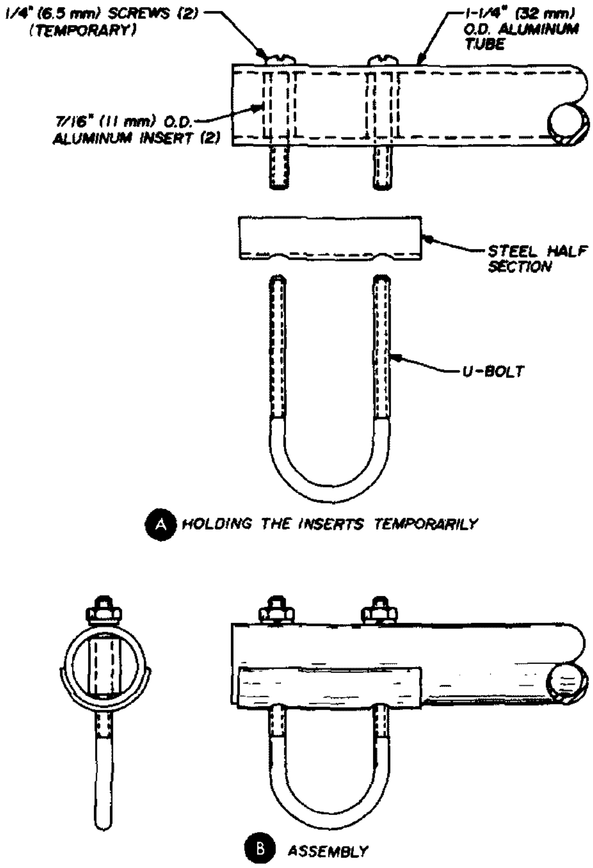


fig. 2. Assembly of the parts for clamping aluminum tubing without deformation. Inserts are held in place temporarily with screws (A) which are pushed out when the U-bolt is inserted (B).

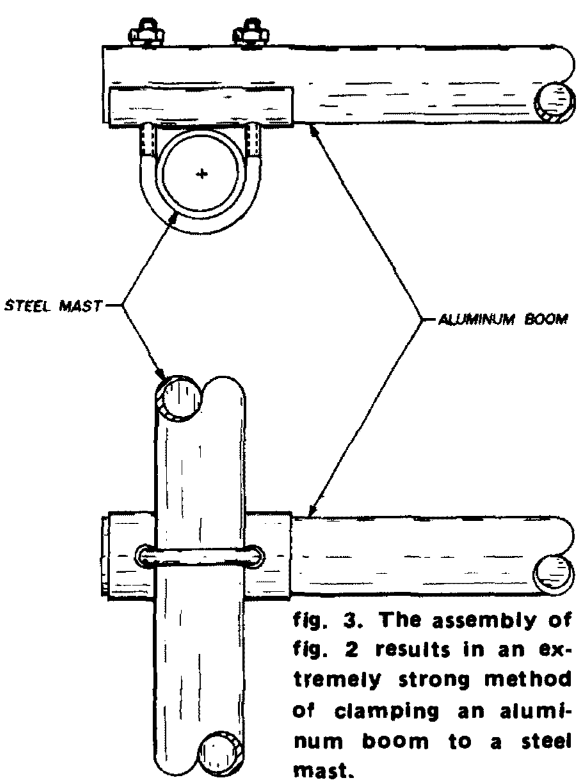


fig. 3. The assembly of fig. 2 results in an extremely strong method of clamping an aluminum boom to a steel mast.

approximately 1-1/8-inch (28.5mm) long, or just long enough to slip crosswise inside the 1¼ inch (32mm) aluminum tube. The finished parts are illustrated in **fig. 1**.

Assembly. It is necessary to insert the two 7/16-inch (11mm) aluminum tubes inside the 1¼-inch (32mm) tubing to mate with the U-bolt holes. The inner tube can be positioned with the aid of long nosed pliers. To hold it in place temporarily, drop a ¼-inch (6.5mm) screw through the hole; then insert the other 7/16-inch (11mm) tube, holding it in a similar manner. The U-bolt can then be inserted from the bottom, pushing the screws out in the process. Before inserting the U-bolt fit one of the steel half-sections on the clamp as shown in the assembly diagram of **fig. 2**.

Using the clamp assembly. Fig. 3 shows a typical installation using the modified clamp assembly. Note that the aluminum inserts and the steel half-tube prevent the 1¼-inch (32mm) aluminum tubing from collapsing under the force of the U-bolt nuts, either in the vicinity of the clamp or at the junction between the two tubes.

Norman J. Foot, WA9HUV

open-wire feeder feedthrough insulator

Open-wire feeders (ladder line or twinlead) should be kept clear of objects which might unbalance the rf currents on the line. Getting the line from indoors to outdoors can thus be a problem. A feedthrough system can be fashioned from about a foot (30cm) of scrap PVC tubing with end plates cut from a plastic bleach bottle (**fig. 4**). Run the line through the wall (a screened attic gable vent makes an easily prepared place) and through the tubing. Cut the end plates with tabs to fit the

tubing. Cut a slot in each plate, and enlarge it at the center so as not to severely bind the feeder. Slide the end plates over the feeder, and then into the ends of the tube.

Although the tabs provide enough tension to hold the plates in place, epoxy or PVC glue will insure they stay put. When the glue is set, slide the tube

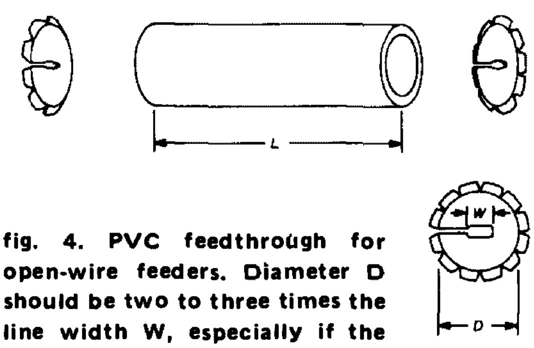


fig. 4. PVC feedthrough for open-wire feeders. Diameter D should be two to three times the line width W, especially if the tube is to pass through metal, solid or screening (the latter being typical of attic vents). The tube should be long enough to allow the feeder line to clear eaves and other objects, both in and out of doors.

into the wall hole. You may want to paint the tube to match the house exterior. Adjust the feeder so that it runs through the center of the tube. A drop or two of epoxy will hold it in place.

Seal the point where the pipe passes through the wall. If the tube passes through screening, hangers can be fashioned from the same bleach bottle used for the end plates. The end plates do double duty by also keeping bugs and squirrels from getting indoors.

L.B. Cebik, W4RNL

dielectric antenna for 10 GHz

Above 1296 MHz it is impractical to use conventional antennas such as simple dipole arrays, Yagis and collinears. Instead, a different class of antennas is found: reflectors (both cylinders and parabolas) and horns. One antenna type which has seen relatively little amateur application is the dielectric rod antenna, probably because few military and commercial stations use them, and secondly, because their design is largely empirical. Finally, the construction of required tapered dielectric sections is beyond most amateur workshop capabilities.

The antenna described in this article, however, can be easily built by the amateur, and consists of a tapered dielectric slab inserted directly into the waveguide. Amateurs using coaxial systems can merely use a coax-to-waveguide adapter. As can be seen in fig. 5, the taper is not smooth, but varies in abrupt jumps every 5cm (2 inches) along the antenna, which is easier to construct. Inside the waveguide, the jumps are every 2cm (3/4 inch). These dimensions are not critical (±0.5cm or ±3/16 inch) but the increments should be uniform. The taper inside the guide matches the impedance of the antenna to the impedance of the waveguide.

The antenna is constructed by laminating strips of 1/10-inch (2.5mm) thick plexiglass window glazing with a solvent

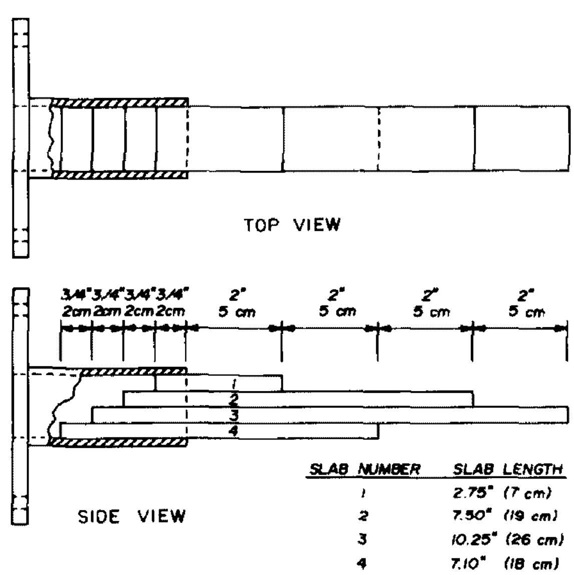
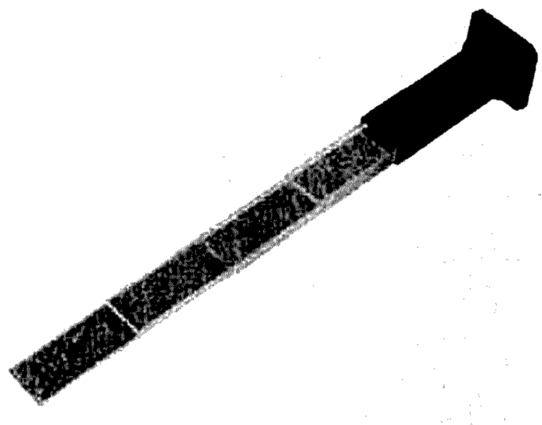


fig. 5. Dimensions of dielectric antenna suitable for use on the amateur 10 GHz band. Vertical and horizontal pattern is shown in fig. 6.



Dielectric antenna for 10 GHz uses laminated strips of 0.1" (2.5mm) plexiglass window glazing. Gain is about 14.5 dB above an isotropic radiator and is flat across the 3cm amateur band.

such as ethylene dichloride. This thickness is chosen because four layers closely match the inside height of the commonly used WR-90 waveguide. All of the strips are 0.9 inch (23mm) wide. The result is a tapered dielectric plug which slides easily into the waveguide. The plug can be secured in the guide with either non-conducting epoxy cement or nylon screws.

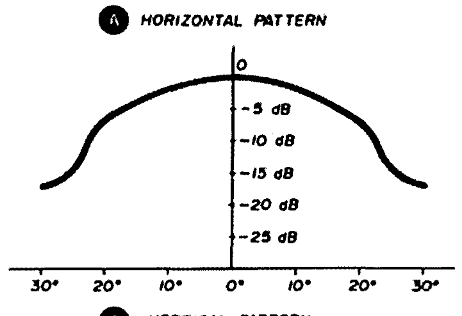


fig. 6. Vertical (A) and horizontal (B) radiation patterns of the dielectric rod antenna at 10 GHz.

A word of caution if you use screws. First, they must not be conductive and second, they should be mounted along the centerline of the wide dimension of the waveguide; otherwise the resultant holes will couple power out of the guide, resulting in lower gain and higher side lobes.

The completed dielectric antenna exhibits a gain of 10 dB above an open-ended section of waveguide or about 14.5 dB above an isotropic radiator, and gain is flat across the 3cm band. The horizontal and vertical radiation patterns are plotted in fig. 6. The vertical beamwidth is 25° whereas the horizontal beamwidth is about 20°.

John M. Franke, WA4WDL

cornell-dubilier ham rotators

Cornell-Dubilier Electronics* has confirmed that the Ham-M and the Ham-II rotators, and also the control units, are interchangeable. Therefore, stand-by spare rotators of either kind can be used with either control unit. Furthermore, all rotator parts have the same stock numbers and internal photographs, so spare parts can be kept on hand and used in either rotator.

In addition to the convenient calibration switch, the addition of a 180-ohm resistor and a 13-volt zener regulator to the Ham-II is a welcome change as it holds the calibration constant under conditions of varying line voltage.

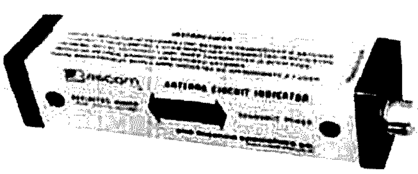
The possibility of burning out the direction indicator potentiometer can be eliminated by placing fuses between the potentiometer and terminal screws 3 and 7, inside the rotator.

Bill Conklin, K6KA

*Cornell Dubilier Electronics, Department C, 118 East Jones Street, Fuquay-Varina, North Carolina 27526.

new products

antenna circuit indicator



A new automatic indicating device designed to warn vhf communications equipment operators of antenna problems has been announced by Ascom Electronic Products, a division of The Antenna Specialists Company. Recognizing the fact that mobile communications antennas are often subjected to some degree of damage, and that the problem is usually not immediately detected by the operator, the antenna circuit indicator immediately alerts the operator to a problem. It also helps prevent damage to equipment by eliminating long periods of transmission into extremely high vswr. The indicator, model ASM-104, is an especially valuable device where maintenance of communications is critical.

The new Ascom antenna circuit indicator is designed for easy mounting under the vehicle dash. Two lights are used to give the operator instant indication of the system's operation. When transmitting, one light indicates the normal condition of rf power being transmitted. Should the second indicator light, showing reflected rf power, it alerts the operator to antenna problems that could affect communications capability. For reliability, LEDs are used for indicators, eliminating bulb burn-out problems. The unit uses no batteries or connection to other power sources, requires only 10 watts or more of rf energy, and operates over a frequency range of 144 to 174 MHz. Complete details on the new ASM-104 indicator are available from Ascom Electronic Products, 12435 Euclid Avenue, Cleveland, Ohio 44106, or by using *check-off* on page 126.

aluminum towers

A new line of fully-assembled, crank-up aluminum communications masts in 40 to 60 foot heights, and prefabricated towers up to 100 feet, has been announced by Fred Franke, Inc. The maintenance-free Aluma Towers are available in a variety of heights and designs for ham radio and commercial and industrial communications systems. The 40-foot tower weighs 56 pounds and the 60-foot model weighs only 95 pounds. Both include the crank-up feature.

The corporation will also manufacture a fixed-guy tower using 10- or 20-foot sections up to a total of 100 feet. The first 50 feet can be assembled on the ground with the remaining units added with special pre-drilled sleeves. The weight of a 10-foot add-on section is only 15 pounds. A special non-guyed tower for mobile homes will also be available. This unit will include a 20-foot high triangular base with a 10-foot extension for a total of 30 feet. The crank-up models, all heliarc welded, are

delivered fully assembled and can be raised on a concrete or driven stake base in less than an hour with no special equipment.

Dealerships for the distribution of the Aluma Towers through retail outlets are being established throughout the country. Inquiries may be directed to Fred Franke, Inc., 1639 Old Dixie Highway, Vero Beach, Florida 32960, or use *check-off* on page 126.

antenna magnetic mount

Using a new type of permanent magnet in a configuration that achieves maximum magnetic pull, Larsen Antennas has just introduced a mobile antenna that has an amazing ability to "stay put." In fact, the hold is so strong that caution must be exercised when placing the mount to be sure you get it where you really want it.

The Larsen Magnetic Mount is available in five different models which accommodate all popular types of mobile antennas including Motorola, GE, DB Products, Antenna Specialists and, of course, all Larsen Models. The unit comes complete with 12 feet (3.7m) of RG-58A/U coax and plug all attached. All that is needed to become operational is to screw the antenna into the base, place the magnetic base on the vehicle roof, fender, trunk or any other convenient spot and run the attached coax to the equipment.

The magnet used on the Larsen mount is guaranteed to be permanent and to stay in place on the vehicle (a flat surface is required) at any highway speed. Design of the mount is such that full capacitance coupling assures adequate ground plane for full antenna efficiency. For more details, write to Larsen Antennas, Post Office Box 1686, Vancouver, Washington 98663, or use *check-off* on page 126.

programmable calculator

A new programmable scientific pocket calculator with significantly more keyboard functions and addressable memories than any previous pocket model was recently introduced by Hewlett-Packard. The new HP-55 is also the first pocket calculator with a built-in digital timer, valuable for short, timed experiments.

The nine-ounce HP-55 features simplified, keystroke programming, with 49 steps of program memory plus branching, testing and editing capability; twenty addressable memories, twice as many as any other pocket model; a total of 86 keyboard functions, operations and conversions; plus a digital timer with 100-hour capacity and the ability to store and recall as many as 10 splits (elapsed time readings within an event). Price of the HP-55 is $395.00 (domestic USA).

Like other HP pocket calculators, the HP-55 features the RPN (Reverse Polish Notation) logic system with a four-memory stack that holds intermediate answers and automatically brings them back when needed in a calculation. Beyond the standard arithmetic, trigonometric and logarithmic functions (including antilog), the HP-55 will work in any of the three trigonometric modes –

More Details? CHECK–OFF Page 126

degrees, radians and grads – and allows the user to convert among any of them. The direct polar/rectangular conversion also enables the user to perform vector arithmetic.

The 20 addressable memory registers in the new calculator permit *register* arithmetic (with 10 memories) and simultaneous two-dimensional vector accumulation. The statistical functions of the HP-55, coupled with the 20 addressable memories, give the user the ability to do two-variable mean and standard deviation, linear regression and linear estimate, curve plotting and four simultaneous linear equations with four unknowns.

The direct conversions on the HP-55, which work both ways, include inches/millimeters, feet/meters, US gallons/liters, pounds mass/kilograms, pounds force/Newtons, degrees *farenheit*/degrees centigrade, and BTU/Joules. Additional capabilities include percentage, root extraction, n factorial (for permutations and combinations), squaring numbers, raising numbers to powers and calculating reciprocals. The constant pi also is included.

HP has added a quartz crystal to the HP-55 circuit to provide a 100-hour timer accurate to ±.01 percent. In the timer mode, the calculator display shows hours, minutes, seconds, tenths and hundredths. While the timer is running, the user can take and store up to 10 splits simply by pushing the digit keys. After the timer is stopped the splits may be recalled by pressing the appropriate digit key.

The HP-55 can display up to ten significant digits with a two-digit exponent and appropriate signs, and comes with an owner's handbook, quick reference guide, program notation pad and an ac adapter/recharger that allows the calculator to be operated on ac while its batteries are recharging. The HP-55, like other HP pocket calculators, will be sold through HP's calculator sales force, by direct mail and through leading college

More Details? CHECK—OFF Page 126

bookstores and department stores.

Two handbooks containing programs for the HP-55 programmable scientific pocket calculator are also available. The \$10.00 books, "HP-55 Mathematics Programs" and "HP-55 Statistics Programs," contain general descriptions of the programs, formulas used in each program solution, numerical examples, user instructions, program listings and register allocations.

For more information, use *check-off* on page 126, or write to Inquiries Manager, Hewlett-Packard Company, 1501 Page Mill Road, Palo Alto, California 94304.

precision balanced mixer

Lithic Systems has announced a precision balanced mixer IC, types LS1596A and LS1596B, which are plug-in replacements for the popular 1596 balanced mixer. The frequency range of the new LS1596A/B has been increased to 250 MHz, and its guaranteed matching characteristics permit untrimmed operation with precision loads and sources. Applications include balanced modulation and demodulation, frequency heterodyning and multiplication, multiplexing and demultiplexing, and phase detection in ssb, dsb, a-m, fm and audio communications systems.

The LS1596A and LS1596B are available in 10-pin TO-100 packages. The LS1596B guarantees 1% internal matching while the LS1596A is specified at 2%. For more information, write to Robert A. Hirschfeld, Lithic Systems, Inc., Post Office Box 478, Saratoga, California 95070, or use *check-off* on page 126.

volt-ohmmeter

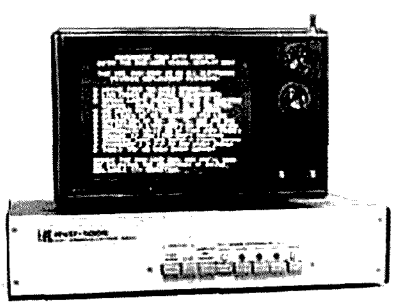
A skillful blending of features, quality parts and performance at an attractive low price may make the new RCA WV-547A volt-ohm-milliameter the best selling general-purpose test instrument

More Details? CHECK—OFF Page 126

in the electronic servicing field, according to RCA Electronic Instrument officials. Features that are not normally found in any vom test instrument selling under $50 are included in the new RCA Tech Vom which is priced at only $19.95. These features include a no-stick taut-band diode-protected meter, dual detent function switch for long life, a rugged high-impact plastic case, accuracies of ±3 percent dc, ±4 percent ac with 20,000 ohms-per-volt dc sensitivity, and five functions with one percent precision resistors on all 19 ranges.

Designed for general purpose testing and servicing applications, the new RCA Tech Vom comes complete with test leads. An optional carrying case (WG-447A) is available for $5.75. Additional information on the RCA WV-547A is available from RCA Electronic Instrument Distributors, from RCA Electronic Instruments, 415 South Fifth Street, Harrison, New Jersey 07029, or by using *check-off* on page 126.

visual display unit



HAL Communications Corporation has announced their new RVD-1005 visual display unit, the second generation version of the HAL RVD-1002, the first visual display unit offered to the teletype operator. The features that have made the RVD-1002 so popular are retained in the RVD-1005, and some new features have been added. For example, the RVD-1005 offers operation at 60, 66, 75 and 100 wpm, manual line feed and letters shift controls, se-

lectable letters-shift-on-space control and automatic carriage return — line feed on line feed (non over-print).

New features include display format of 25 lines, 40 characters per line; automatic carriage return; and speed indicator. The automatic carriage return will line feed on space when space is received after the 34th character. Prevents splitting of short words (will not split words of six characters or less). The speed indicator times the incoming signal and causes an LED to light, showing the operator which speed switch to select. This is especially helpful when tuning unknown commercial stations.

Operation of RVD-1005 is similar to that of the RVD-1002. The incoming signal from the terminal unit, or keyboard, is decoded, stored, and a video signal is generated which is fed to a video monitor or modified TV set. The TV set modification consists only of coupling the video through a capacitor to the first video stage of the set. The TV set should have a power transformer so the chassis is isolated from the power line.

The price of the RVD-1005 is being held at $575, postpaid in the continental USA (except California, Oregon and Washington, add $5). For more information, write to HAL Communications Corporation, Box 365, Urbana, Illinois 61801, or use *check-off* on page 126.

wireless telegraphy

For many years, beginning around 1800, the Royal Institution in London sponsored a series of lectures on subjects related to the physical sciences, astronomy and geology. In 1851, when the regular publication of abstracts began, the Institution had already been a major research center for fifty years. Halstead Press, in their *Royal Institution Library of Science*, is reproducing many of the more important lectures. One of the volumes in the series, *Wire-*

More Details? CHECK—OFF Page 126

less Telegraphy, Edited by Sir Eric Eastwood, traces the development of telecommunications from the 1850s, when the submarine telegraph and Atlantic cable were first discussed. Attention turned to telephony in the 1870s, and then to "Signalling through Space Without Wires," a famous lecture by William Henry Preece in 1897. From 1900 on the lectures were devoted to the development of wireless telegraphy, radio telephony and microwaves.

Marconi was a member of the Royal Institution and a frequent contributor to its famous series of Friday night discoures. Many of his lectures are contained in this book including "Wireless Telegraphy" (1900), "Transatlantic Wireless Telegraphy" (1908), and "Radio Communications by Means of Very Short Waves" (1932). John Ambrose Fleming, most famous for his invention of the diode detector, lectured on developments in radio communications and thermionic vacuum tubes.

Amateurs who are interested in the history of telecommunications will find this series of twenty-two lectures, given over a period of seventy-five years, to be both entertaining and educational. 391 pages, hardbound, $32.50 from Halstead Press, 605 Third Avenue, New York, New York 10016.

programmable operational amplifier

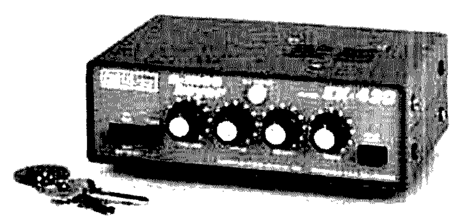
A new programmable operational amplifier which permits tailoring the electrical parameters to the user's needs has been introduced by Motorola Semiconductor. Programming is achieved by a choice of an external resistor value of a current source applied to the I_{set} input. This allows optimization of dc characteristics such as input current, power consumption and bias current, and ac characteristics such as open-loop voltage gain, slew rate and gain-

bandwidth product, to the user's requirements.

The MC3476 operational amplifier is designed to operate over a power-supply voltage range from ±6 volts to ±15 volts. Its low power consumption of 4.8 mW (typical) makes it particularly useful in battery operated equipment. Other features include low input offset and bias current (a maximum of 25 and 50 nanoamps, respectively) and a high input resistance of 5 megohms, typical. The amplifier requires no frequency compensation and has offset null capability and short-circuit protection.

Despite electrical characteristics which compare favorably with some of the better op-amp specifications, the MC3476 operational amplifier price is competitive with some of the lowest cost units currently available. The device is available in both plastic and metal packages. For further information, contact the Technical Information Center, Motorola, Inc., Semiconductor Products Division, P.O. Box 20924, Phoenix, Arizona 85036, or use *check-off* on page 126.

cmos electronic key



A flea power cmos electronic keyer just announced by Curtis Electro Devices uses the new 8043 IC keyer to reduce complexity and increase reliability. The EK-430 contains all the most desirable keyer features such as self-completing dots, dashes and spaces, instant starting clock, iambic operation, dot memory, element weight control

More Details? CHECK—OFF Page 126

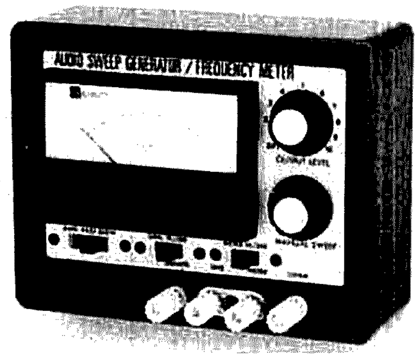
and built-in sidetone with speaker. Powered by either 117 Vac or an inexpensive 9-volt battery, transistor output switching keys up to plus or minus 300 Vdc at 200 mA maximum. Key debouncing circuitry assures trouble free operation on any type of key paddle regardless of the condition of the contacts.

Housed in a low profile, two-tone blue metal cabinet weighing only 26 ounces, the EK-430 is priced at $124.95 fob the factory. For further information, write to Curtis Electro Devices, Inc., Box 4090, Mountain View, California 94040, or use *check-off* on page 126.

linear ic principles, experiments and projects

This new book by Ed Noll, W3FQJ, was written to introduce the principles of operation of the integrated circuit. Chapter 1 covers basic semiconductor principles: basic IC structures, the pn junction, the bipolar transistor, transistor fabrication and the field-effect transistor (FET). Succeeding chapters explain integrated-circuit structures, basic circuits, operational amplifiers, multipurpose and special ICs. The last four chapters give a broad coverage of how linear ICs are used in commercial, industrial and test equipment. Home-entertainment audio, a-m, fm and television applications are also stressed. Amateur radio and shortwave enthusiasts will find the projects in Chapters 9 and 10 good examples of the many opportunities for the use of ICs in the two-way radio field. These final chapters also will be appreciated by those who like the learn-by-doing approach; they'll find an intriguing collection of school, lab and home construction projects. 384 pages, softbound, $8.95 from Ham Radio Books, Greenville, New Hampshire 03048.

audio sweep generator and frequency meter model 140B



The model 140B is virtually three instruments in one: an audio oscillator, audio sweep generator and frequency meter. It may be operated manually as a conventional oscillator, or swept automatically as a voltage-controlled generator. The generator sweeps over two ranges: a high range from 1 kHz to 20 kHz, and a special expanded low range from 40 Hz to 1 kHz, which allows for a close look at low-end frequency-response problems.

The frequency meter may be used independently. However, when connected to the output of the sweep generator the frequency reading is continuously displayed. Unlike digital counters, which are limited to measuring fixed frequencies, this frequency meter may be used to monitor the changing frequency of the sweep generator or any other external frequency source.

The model 140B is an excellent input source for testing amplifier frequency response, speaker enclosures and tape recorder head alignment. This low cost instrument provides a sine wave output that is variable from 0 to 2.5 volts p-p. Once set, the amplitude remains flat over the entire frequency range. (±¼ dB) sine wave distortion is less than 1.5%. Square wave output is fixed at 8 volts p-p. $78.95 from Production Devices, 7857 Raytheon Road, San Diego, California 92111. For more information, use *check-off* on page 126.

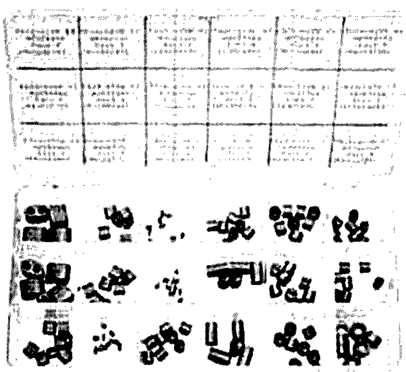
More Details? CHECK—OFF Page 126

semiconductor cross reference

The new edition of the semiconductor cross reference and transistor data book is now available from International Rectifier Corporation. The 70-page brochure lists over 44,000 parts and corresponding IR replacements including rectifiers, diodes, zeners, transistors, SCRs and ICs. All are indexed in straight alpha-numeric sequence to facilitate location of the desired part number. Also included are four pages of transistor specifications, showing polarity, case style, maximum current and typical bandwidth and gain. A data sheet on IC replacements lists absolute maximum rating and case style.

To obtain a copy of the cross reference (JD601), write Semiconductor Division, International Rectifier Corporation, 233 Kansas Street, El Segundo, California 90245 or use *check-off* on page 126.

ferrite-core kit



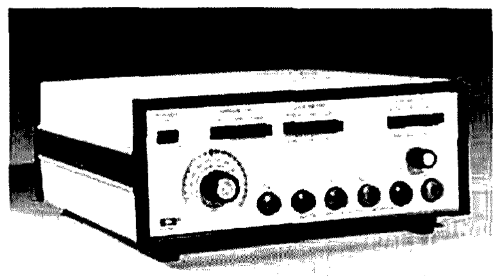
A sample kit containing 18 nickel-zinc and new high permeability manganese-zinc ferrites for winding wideband transformers including baluns can be obtained for $10.00 from the Fair-Rite Corporation. These cores are useful over a wide frequency range but were selected for their effectiveness at frequencies above 2 MHz. Applications include radio communications, vhf receivers, mobile communications and

other systems operating at high frequencies.

The 18 ferrite core shapes are provided in three different materials along with application data including curves of parallel impedance, reactance and resistance per turn of wire for each sample core. Data also describes how to choose the right material and core size for a particular application.

Application data is available without charge. The sample kit can be ordered by sending a check or purchase order for $10.00 to Fair-Rite Corporation, Wallkill, New York 12589. For more information, use *check-off* on page 126.

sweep/function generator



A new low-cost sweep/function generator oscillator has been introduced by Exact Electronics. The model 195, priced at $149.50, is expected to make inroads into audio testing and amateur electronics applications where high performance function generators were not previously available. The new instrument, housed in a compact, rugged case, produces sine, square, triangle and swept waveforms as well as fixed amplitude pulses. Its frequency range is from 2 Hz to 200 kHz in three ranges with a linear/logarithmic frequency control.

An internal sweep generator will sweep 1000:1 (3 decades) on any of the three main frequency ranges. This permits technicians to sweep, either linearly or logarithmically, the entire audio range of amplifiers or speakers without changing ranges or even turning a knob. High and low level sine outputs with

More Details? CHECK—OFF Page 126

amplitude control of both are provided. A voltage control frequency (VCF) input permits controlling frequency from an external source. Fully portable, the model 195 is powered by a 9-volt transistor battery, completely eliminating any problems with 60-Hz hum. An optional rechargeable power supply is available.

For more information, write to Exact Electronics, Box 160, Hillsboro, Oregon 97123, or use *check-off* on page 126.

solder slide rule

A handy new slide rule published by Kester Solder provides comprehensive flux selector data on one side and offers solder alloys guides on its flip-side. The pocket size Kester slide rule gives flux choices for 22 metals, ranging from easy-to-solder gold, copper, and tin to very-difficult-to-solder chromium and stainless steel. Thirty-six solder alloys also are listed, along with temperatures at which the solder becomes plastic and becomes liquid. Readers may obtain this slide rule by writing to Kester Solder, 4201 Wrightwood Avenue, Chicago, Illinois, 60639, Attn: Mack Haraburd.

touch-tone decoder board

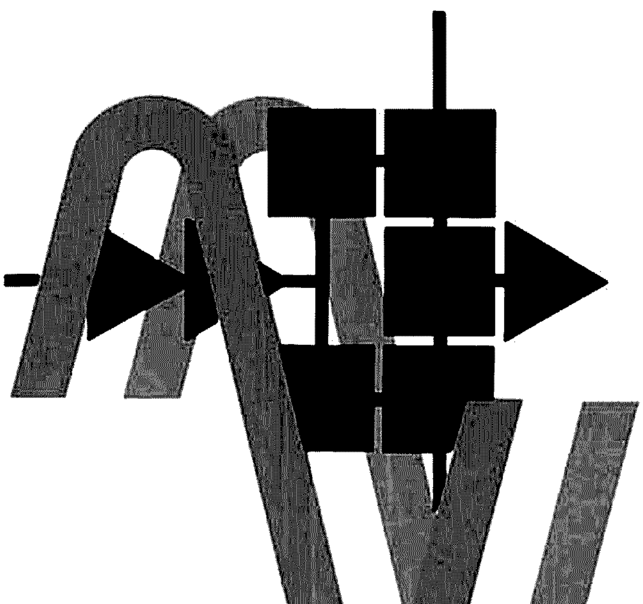
A new seven-digit *Touch-Tone* Decoder board kit with multiple capabilities has been announced by CTI Manufacturing Company. The new board features full seven-digit Touch-Tone decoding and encoding, uses cmos circuitry, and is installed on a miniature 2¼x3¼-inch (57x82mm) circuit board. The board can be mated with an optional Touch-Tone pad, if desired, or used with existing equipment. The basic kit for the decoder board is priced at $59.50. For more information, write to CTI Manufacturing Co., Post Office Box 1422, Corinth, Mississippi 38834, or use *check-off* on page 126.

$1.00

focus on communications technology . . .

ham radio
magazine

JUNE 1975



A PHASING-TYPE SINGLE-SIDEBAND TRANSMITTER

this month

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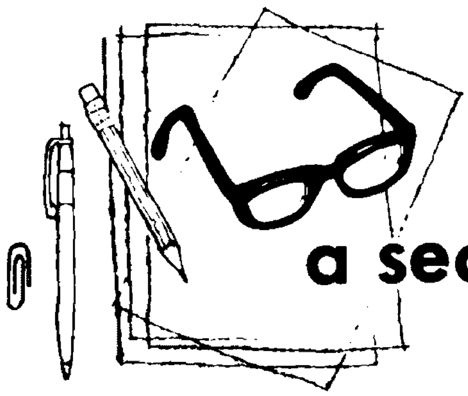


a second look

by jim fisk

As I pointed out in this space last month, the FCC is considering various new ways of determining the power limitations of amateur transmitters. These range from dc power input and power output to manufacturer's plate dissipation ratings. All of these methods, however, contain one or more variables which are subject to interpretation and policing. Dc input measurements, for example, must be done with calibrated meters at the time of operation. Output power measurements are complicated, and require accurate instrumentation. Plate dissipation ratings are aribtrary numbers established by tube manufacturers which are based upon a given amount of air and back pressure to establish a desired plate dissipation rating. However, it is conceivable that a manufacturer could rate a large tube at a *lower* plate dissipation rating, or conversely, by requiring more air flow, considerably *increase* the dissipation capability of the tube. This variation is brought about by the fact that different types of service are used in communication systems (class A, B or C); flexible data ratings for different classes of service must be established because of the variations in plate efficiency from one class of operation to another.

As has been pointed out by Jack Quinn, W6MZ, of EIMAC, there is only *one* common denominator in a vacuum tube which determines the maximum capability of that device, and that is the manufacturer's rated heater, or filament, power. This is a parameter which is very carefully established by all tube manufacturers and follows rigid, fixed laws of physics. These current/voltage relationships of the emitter cannot be increased with any degree of freedom without suffering short tube life or catastrophic failure.

Rather than establishing new amateur power limitations based upon plate dissipation ratings or publishing a list of approved tube types which must be continually maintained and up-dated, Quinn has proposed that certain *maximum filament or cathode heater power ratings* be established. For example, Extra, Experimenter and Advanced class licensees could use a final amplifier having one or more thoriated-tungsten filament tubes with a total filament power rating which does not exceed 200 watts. An amplifier with an indirectly-heated oxide cathode tube would have a total heater power which does not exceed 60 watts, according to the manufacturer's ratings. If this were done, the ratings would be based upon a common ground and good, sound, technical background. This would also be compatible with amateur equipment in common use today as shown in table 1.

Power levels for the General-Technician and Communicator-Novice classes could be scaled down by whatever percentage the Commission deems

(continued on page 60)

presstop

ARRL FORUM AT DAYTON HAMVENTION in late April, led by League president W2TUK, was primarily a question and answer session. As expected, restructuring was the primary subject but, since the League position isn't officially set until the Director's meeting in May, Harry's answers had to be his own personal opinions. Other subjects included possible ARRL insurance program (under active consideration but hard to find an underwriter to cover all states), how to communicate with the FCC, protecting the ham bands from incursions by other services, and type acceptance of commercially made gear.

ARRL 20282 Survey has had an outstanding response — about 56,000 of the League's 100,000 members returned their surveys in time to be processed!

RESPONSE TO RESTRUCTURING so far received by FCC summarized by Prose Walker at Dayton. Two-thirds support 20282, one-third oppose. Most want 10 meters shared. Of the 80% discussing power, more than half are against reductions to Generals and Techs and all oppose levels offered Novice/Communicator. Measuring output power is opposed by the majority.

TYPE ACCEPTANCE of amateur gear is becoming more and more of an active issue at the FCC. The subject was discussed at the Dayton Hamvention Novice forum, which featured both Charles Higginbotham, Chief of the FCC's Safety and Special Services Bureau, and Ray Spence, W4QZW, Chief Engineer of the FCC.

Type Acceptance has some good features but some bad ones, too. It might clean up some problem-causing commercial gear and thus help reduce the RFI problem. At the same time it could prove very stifling if, for example, it forbade the individual amateur from modifying his equipment to meet his own needs.

PROPOSED NEW HF HAM BANDS won't be ours without a stiff fight in 1979. North Atlantic broadcasters would like to increase the international BC allocations between 3 and 27 MHz by another 7 MHz at the forthcoming world conference! For reference, that's more than twice what they're presently allocated.

NEW MOTOROLA COMMERCIAL FM GEAR is pushing the amateur price range — Maxar line of 150- and 450-MHz transceivers starts at $395 for a 10-watt out model with several channels. For repeaters their new Spectra-Tac voting system looks very competitive, and finally, they have a new pager receiver which shakes in your pocket when you get a call.

JA1ANG VISITED AMSAT early last week and reported JAMSAT is well along in construction of a two-meter-to-435-MHz transponder in anticipation of the next OSCAR. Other JAMSAT projects are also in the mill.

TI9DX On Cocos Island was heard on OSCAR recently, and TU2EF reports he's been quite active on Mode A — his log would make a serious 20-meter DXer drool. FY7AS, A2CJP, 4W1ED and ZS3E are samples — he reports EA8CS may also be on soon.

"Area Coordinator" is the tentative title for a position established by the AMSAT board. Purpose for the Area Coordinator, who should be an active OSCAR user, is to advise amateurs interested in getting on OSCAR and provide local clearinghouse for OSCAR activities. Active satellite users who want to help out in this capacity should contact AMSAT headquarters.

FCC TAKING A CRITICAL LOOK AT SPEECH PROCESSING if footnote at the end of Docket 20282 is any indication, but no response to the restructuring docket filed so far has even acknowledged that paragraph's existence. Since DXers are the most frequent users of processors, we better take another look at that paragraph and its implications — mushy and broad signals have all too often caused problems on the lower ends of the DX bands.

MORE NON-AMATEUR OPERATIONS THREATEN 420-450 MHz band, as TI in Dallas attempts to get a new Special Temporary Authority to continue operating a high-powered navigation system on 430 MHz. The 20 kW ERENS (Extended Range Electromagnetic Navigational System) transmitter has a range of 250 miles and has been on since last September. They have also filed requests for similar systems on Cape Cod and Montauk Point, Long Island. If permitted, these pulsed navigational systems would make a large portion of the 420-450 MHz band practically unusable.

phasing-type single-sideband transmitter

A direct-conversion companion transmitter for the Phase II ssb receiver

G.K. Shubert, WAØJYK, 1308 Leeview Drive, Olathe, Kansas 66061

The Phase II ssb receiver described last year in *ham radio*[1] has been a moderate success and has proven worthy of a companion transmitter. The phasing method of single-sideband detection has been revived slightly, and the addition of a good ssb transmitter will even stir interest among the "filter freaks." After all, doesn't it make sense to generate just one sideband with phasing techniques rather than to generate both sidebands and remove the unwanted one with a brute-force crystal filter?

This article presents the Phase II reciter, a ssb exciter that borrows phase quadrature rf from the Phase II receiver. A 300-watt PEP solid state linear power amplifier will be the subject of a future article and will round out the Phase II system including vfo, transmit/receive switching, vox, filtering, and a regulated, low-voltage, high-current power supply.

Readers of the Phase II receiver article have indicated an interest in a more elaborate receiver with additional frequency coverage. New circuits for improving the receiver and ideas for future development will be discussed. The possibilities for the receiver and transmitter pair are far too numerous to present in full detail, so individual circuits and interconnect information will be given, leaving the final enclosure and layout to the individual builder. Layout might vary drastically between a receiver-transmitter for a single, crystal-controlled frequency, and one designed for the three lower amateur bands with vfo control.

ssb reciter

Fig. 1 is the classic diagram of a phasing-type ssb transmitter that has appeared in the electronic handbooks for three decades.[2] The 90-degree rf phase is developed in the Phase II receiver and it will be necessary to refer

to the receiver article for additional information on this aspect of the circuit. Since the transmitter may be operated independently if this portion of the receiver is duplicated in the transmitter, this oscillator and phase-shift circuit are described later in the article.

Much of the research done on phasing circuits in the 1940s occurred during the development of telephone carrier systems and many of the schematics bear the Western Electric title box. Engineering texts tell us that a single-sideband-suppressed-carrier signal can be regarded as the resultant of quadrature modulation of a carrier by a pair of signals in phase quadrature, further explained with some pretty involved mathematics.[3] It is possible to switch the phase-quadrature rf to in-phase rf and generate phase-modulated signals. Perhaps this is why the phasing technique was popular in the late 1940s and early 1950s when narrow-band fm was far more popular on the high-frequency bands than the then newer mode, single sideband.

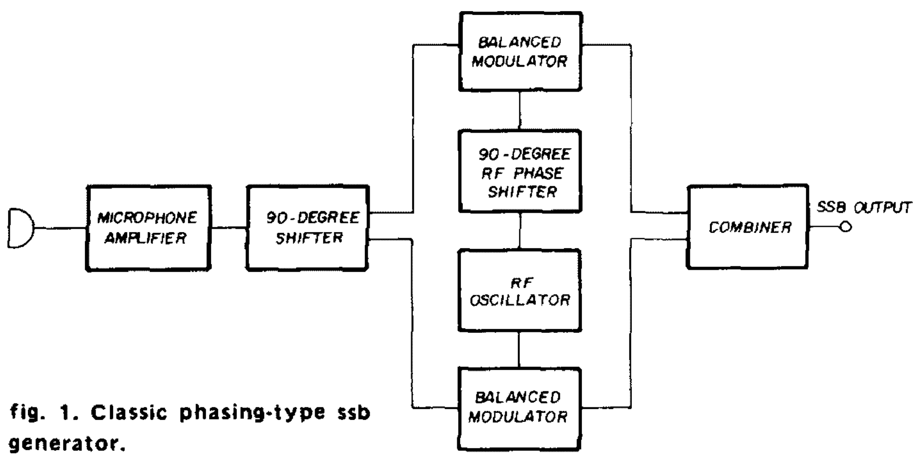


fig. 1. Classic phasing-type ssb generator.

Fig. 2 is a block diagram of the entire Phase II system. The digital phase-shifter in the receiver is used to supply phase-quadrature rf to the exciter. This means the parts cost for the exciter will be considerably less than for the receiver. The exciter uses exactly the same audio lowpass filter and audio phase-shift network as the receiver. It is better to duplicate these components in the interest of simplicity of switching and balancing circuits.

Although these components are not particularly expensive, they may be a little difficult to locate. The 1% precision film resistors are standard MIL-BEL values as are the 0.1-μF capacitors. The 0.028-μF capacitors may have to be made up by paralleling a 0.027-μF and a 1000-pF silvered-mica capacitor. Remember that ultimate sideband suppression depends upon the accuracy of these components, so be as accurate as possible.

A circuit board has not been seriously considered for several reasons. One very obvious reason is that a circuit board takes time to develop and debug, not to mention the problems encountered when you try to make modifications. The exciter is much simpler than the receiver and has fewer ICs (the digital rf phase shifter is the complicated part and it is already on the receiver board). Even the regulated 10-volt supply of the receiver may be tapped. However, if the 10-volt receiver supply is used, a larger heatsink is necessary and a heavier transistor may be needed.

A Motorola MC7812 or Fairchild

μA7812 can be used to replace the 10-volt supply in the receiver and will supply all the 12-volt current (up to one amp) that is required in both the receiver and exciter. However, a good 15- to 24-volt dc supply will be required to maintain good regulation with this voltage-regulator IC. Mobile operation with a 12-volt supply would be impossible with this IC, and the MC7808 voltage-regulator IC is a little too low.

There is sufficient room left on the 5 x 6-inch (12.7 x 15.2-cm) board (same as the receiver board) for vox or a small 10- to 15-watt PEP linear amplifier.* Remember that a transmitter needs a good, heavy, continuous ground plane under *all* the components. Preserving maximum ground area helps isolate the receiver from the transmitter as well as keeping the transmitter calm. The receiver is actually operating all the time but is muted during periods of transmission.

One very versatile method of making permanent breadboards is to use double-sided PC board material with push-in Teflon terminals which can occasionally be purchased on the surplus market. Press-in terminals are about twenty cents each when purchased in small quantities so you may want to use a circle cutter to isolate small islands from the board or use small pieces of PC board material cemented to the main board with a hot-melt glue gun. Layout is not particularly critical except for location of the mixers as detailed later. The large copper area provides both grounding and shielding.

*A printed-circuit board for the Phase II receiver is now available from D.L. McClaren, W8URX, 19721 Maplewood Avenue, Cleveland, Ohio 44135, for $10.50. The transformers for the receiver are still available from the author, WAØJYK (the same transformers are used in the Phase II reciter). The price will remain at $10 for the set of three required for the receiver; the pair required for the reciter is priced at $8. A complete set of five transformers is available for $16.00. This price will be valid only until the present supply is exhausted — new supplies will probably be priced 10% higher to meet the demands of inflation.

circuit operation

The schematic for the reciter is shown in **fig. 3**. The microphone amplifier is a high-gain, high-impedance stage followed by an emitter follower which provides 1000-ohm output impedance to drive the audio lowpass filter. The audio lowpass filter is identical to the one used in the receiver. It is possible to get by with a two-coil filter in the transmitter but the design from the receiver was convenient and provides a little better performance. If a different filter is desired it can be designed from the available literature.[5,6] If tradeoffs are to be made in the filter, or an improvement made, the outcome can be predicted with the help of graphs.[7] Selective bypassing of the audio amplifiers also helps to roll-off the low audio frequencies. The ferrite pot-core assemblies have metal frames and are better shielded from rf fields than the toroids. If power in excess of 100 watts is contemplated, a metal rf shield may be needed to cover the 88-mH coils.

The audio agc system uses a Motorola MC1590G or the less expensive consumer products counterpart, the MC1350P. This versatile gain-controlled amplifier has its internal circuitry revealed in **fig. 4**. For layout purposes, pin 1 is interchangeable with pin 3 (pin 6 with pin 4 of the MC1350) and pin 5 may be interchanged with pin 6 (pin 1 with pin 8 of the MC1350). These terminals are the differential input and output of the amplifier. In the reciter it is being used as a single-ended amplifier because it is easier to keep it stable and maximum available power gain is not required.

The detector for this gain-controlled amplifier is at the output of the exciter where the rf power level is about 1-watt PEP. Detector drive could be obtained from a higher level stage with a voltage

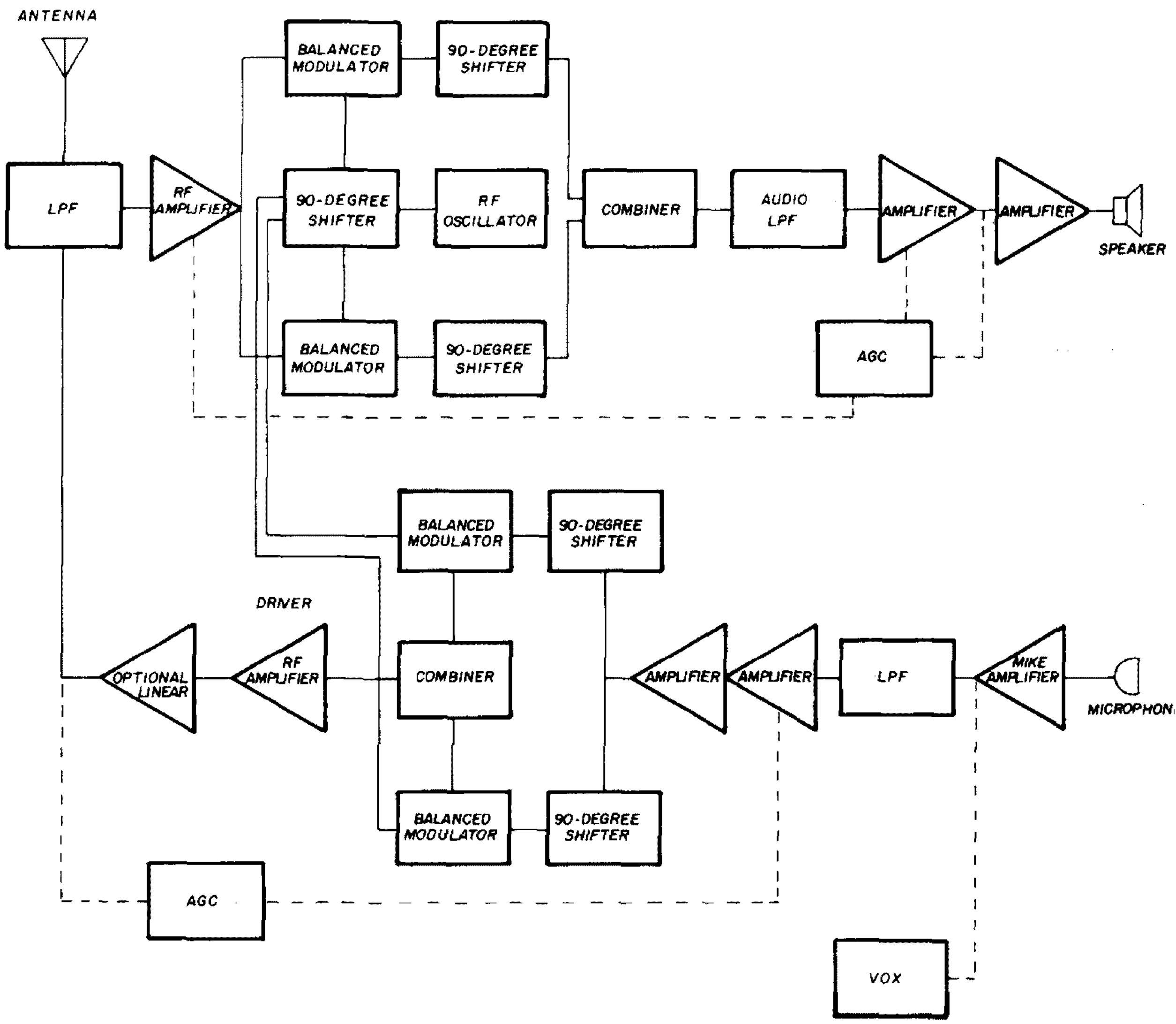


fig. 2. The Phase II ssb transceiver consists of phasing-type receiver, above, and phasing-type ssb generator, below. An optional linear amplifier is discussed in the text.

divider arrangement but it is not desirable to have too much gain in the agc loop. Any imbalance in the balanced modulators will cause some carrier to be present and will be detected by the agc detector. This carrier is interpreted by the detector to be a ssb signal caused by voice modulation and it will promptly turn down the modulation level which will, of course, reduce the level of ssb to carrier and further aggravate the situation.

It is possible to transmit compatible a-m by unbalancing one balanced modulator and sending one sideband with as much carrier injected as is desired. Full a-m can be transmitted by disabling one balanced modulator entirely and unbalancing the other. The combination of audio shaping with the lowpass filter and low-frequency rolloff caused by the audio transformers and selective bypassing (below 300 Hz) with audio compression caused by the agc system makes an effective built-in speech processor. There is no need for any additional speech processing since the percentage of average-to-peak power is quite high already.

The output of the agc-controlled amplifier is fed to another amplifier which has enough power capability to

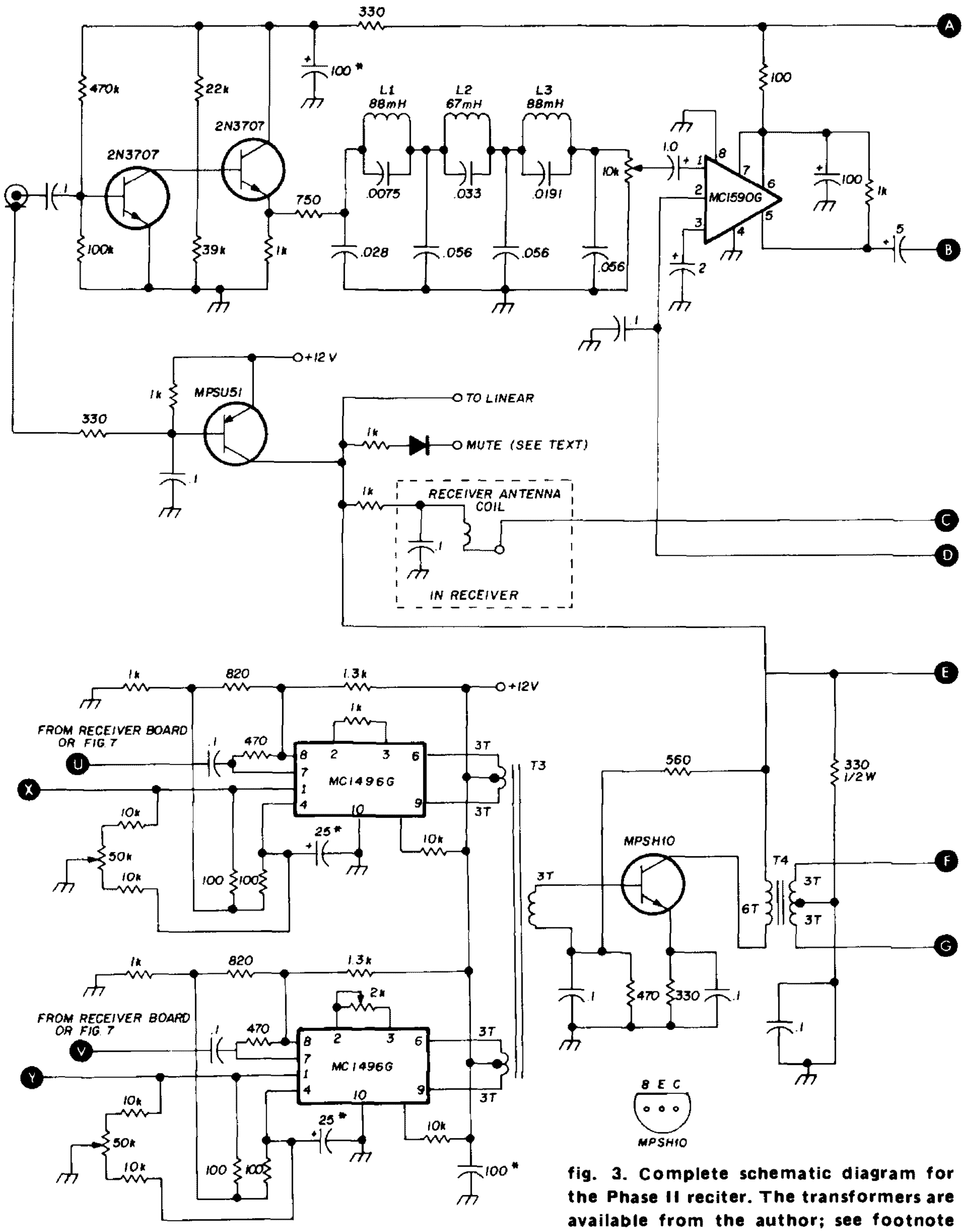
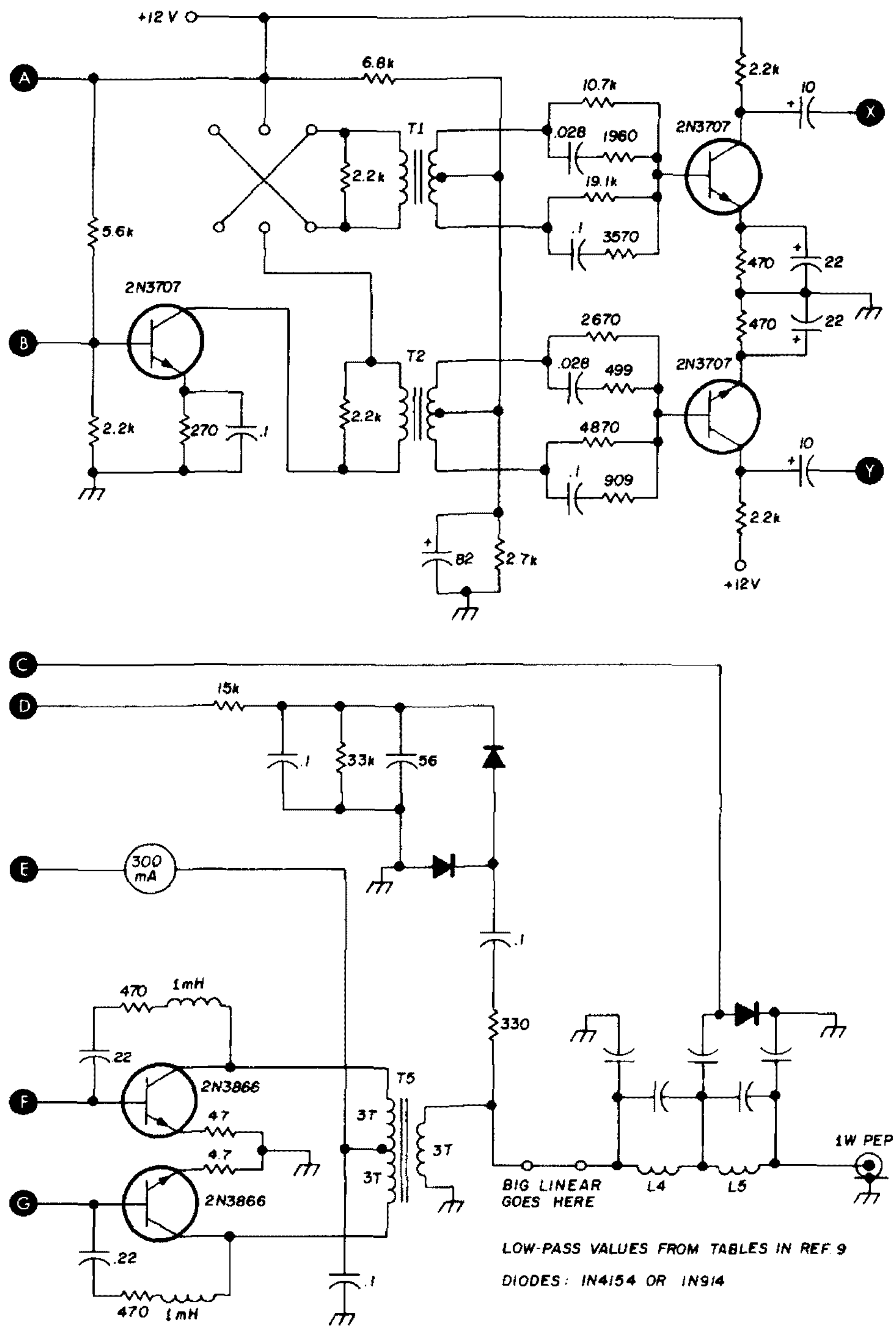


fig. 3. Complete schematic diagram for the Phase II reciter. The transformers are available from the author; see footnote on page 10.

drive the transformers and audio phase-shift networks. It might be possible to directly drive the transformers from the MC1590G ICs if special transformers were designed to do the job, but in the interest of saving time I used the same transformers that were used in the receiver.

The audio phase-shift network is also identical to the one in the receiver and it must work into a relatively high impedance. The input impedance of the



balanced mixers is low, on the order of 200 ohms (set by the bias resistors), so a low-gain buffer stage is necessary between the audio phase-shifter and the balanced modulators. The reversing switch at the primary of one of the audio transformers allows selection of the desired sideband. The switch may be remotely mounted on a panel or the edge of the board with twisted leads since the level at this point is relatively high. The receiver sideband selector switch doesn't offer this option because of the extremely low audio levels.

balanced modulators

The balanced modulators are a pair of Motorola MC1496G integrated circuits. The National LM1496H or Fairchild μA796HC may also be used. If

better performance over a wider temperature range is desired, the Motorola MC1596G or National LM1596H can be used but other components should also be of premium quality. The plastic cased version of this IC should be avoided in this application. The balanced modulators should be located directly under U6 in the receiver so that 0.1-μF disc ceramic capacitors can be connected from pins 2 and 14 of U6 in the receiver, through feedthrough terminals on the exciter board, and terminating at pins 7 (or pin 8) on each of the balanced modulators.

Fig. 5 shows the internal schematic of the MC1496 which presents some possibilities for a simplified board layout. There was some confusion about the MC1496 product detectors used in the receiver because pins 1 and 4 were interchanged on the two detectors to aid in board layout. Careful inspection of **fig.** 5 shows that pin 1 can be interchanged with pin 4 and pin 6 may be interchanged with pin 9 because these are the inputs and outputs of differential stages. This can be quite a help when laying out a PC board or making a breadboard and trying to minimize the number of crossover leads. This transposition of pins is similar to that mentioned for the MC1590G (or MC1350) earlier. The MC1496 is difficult enough to bias for single supply operation because of the 11 resistors, so this trick is well worth remembering.

The 50k carrier-balance pots may be run to the edge of the board for convenience. For ease of adjustment it's a good idea to use multi-turn units. If the balance pots are mounted on a panel the leads should be bypassed for rf. The metal can of the MC1496G is internally connected to pin 10 which is grounded but it is helpful to run a piece of number-18 or -16 wire across the tops of both ICs and solder it to the cans and to the board. This forms a very stiff ground plane between the input and output of the balanced modulators. Carrier suppression and noise improved 6 dB with the extra ground strap. The MC1496G is a very well balanced mixer and feed-around becomes greater than actual feedthrough.

The outputs of the two balanced modulators are summed in the rf transformer T3, an extremely broadband ferrite-core transformer. The broadband coupling eliminates any tuneup procedure but presents some other problems. The balanced modulators will suppress the carrier by 60 dB or so, but the second harmonic of the carrier is suppressed only about 30 dB. The exact amount of suppression will vary with individual units as well as with temperature, layout and frequency.

This second harmonic poses no particular threat but does point out the need for the lowpass filter following the exciter in addition to a lowpass filter after the final linear power amplifier. If the second harmonic of the carrier went unsuppressed to the antenna it would only amount to a few milliwatts but if it were allowed to pass through the solid-state linear power amplifier it would come out at about a quarter watt and, due to the agc action in the linear itself, when there was no voice modulation it could go to two or three watts.

The ferrite core used for transformers T3, T4 and T5 is the Ferronics 12-360J, an ideal configuration for small receiver and transmitter coils in the high-frequency range. Several of these cores are also used in the solid-state linear power amplifier. These two-hole ferrite beads may be wound with all leads coming out one end or with leads coming out opposite ends, depending on which is more convenient for mounting and conserving board space.

linear amplifier

Transformer T3 drives a class-A amplifier stage. A number of transistors will work in this stage but the Motorola

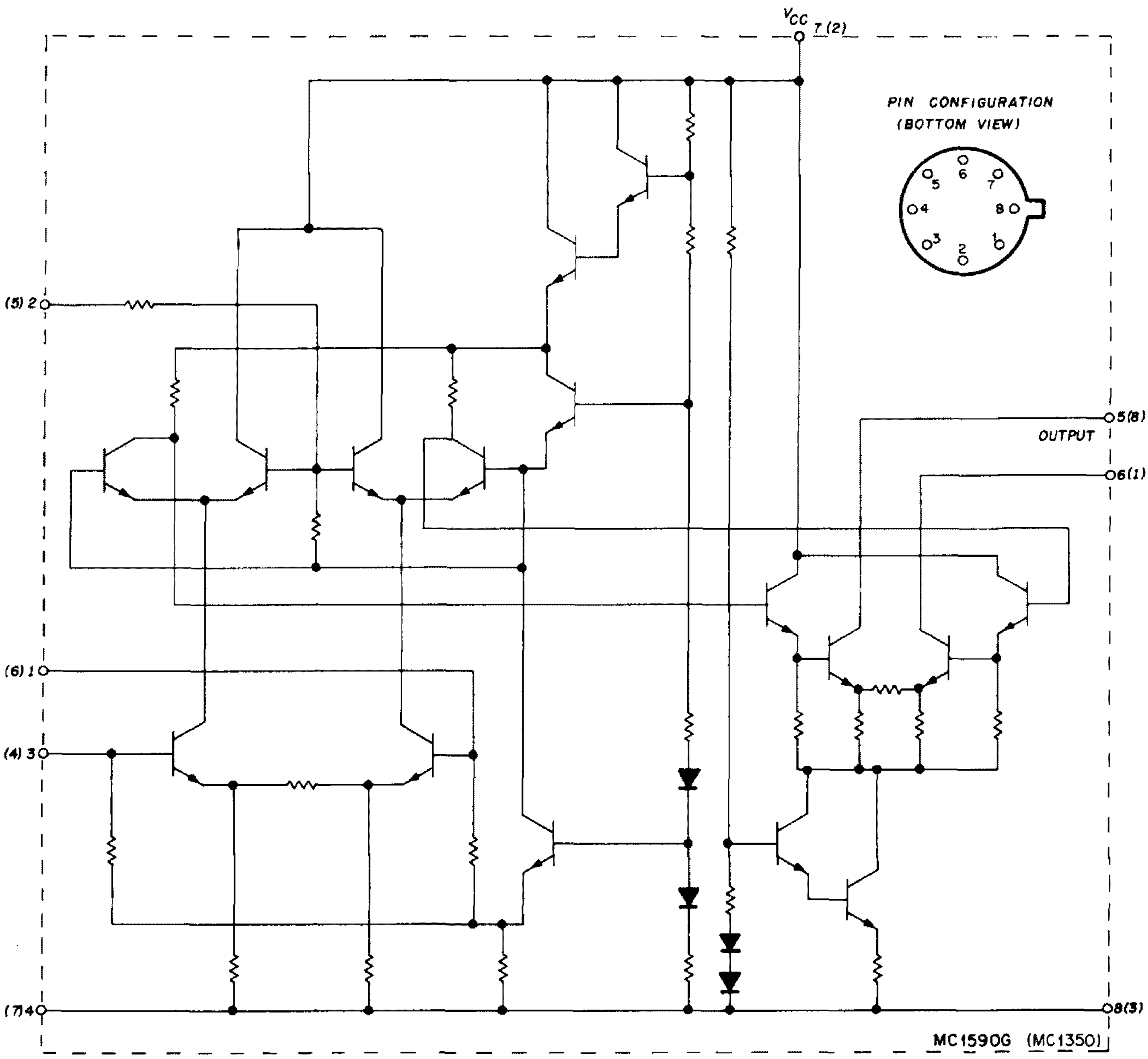


fig. 4. Schematic diagram of the MC1590G (MC1350P) gain-controlled amplifiers. Pin numbers for the MC1350P are shown in parenthesis.

MPSH-10 is economical and performs quite well. A 2N3904 will work too. The base configuration of the MPSH-10 is unconventional and is shown in **fig. 3**.

The output stage is operated push-pull to cancel the even-order harmonics and is transformer-coupled with broadband transformers. This push-pull stage is biased for pseudo-class-AB2 operation since class-B will result in too much distortion and class-A is too inefficient. The 330-ohm bias resistor may have to be adjusted over the range from 220 to 470 ohms so that the final collector resting current is between 10 and 20 mA. The final collector current will rise to 100 mA on voice modulation.

The final transistors are 2N3866s and are occasionally available on the surplus market for under a dollar. The 2N3866 is rated to frequencies as high as 500 MHz and many of them have an actual f_T of 1000 MHz, so many of the units that are factory rejects because of low

f_T or low current gain are usable in this circuit. The emitter resistors are not bypassed and parasitic suppression networks are used between the base and collector leads. More power can be coaxed from the finals by bypassing the emitter leads but there is no protection

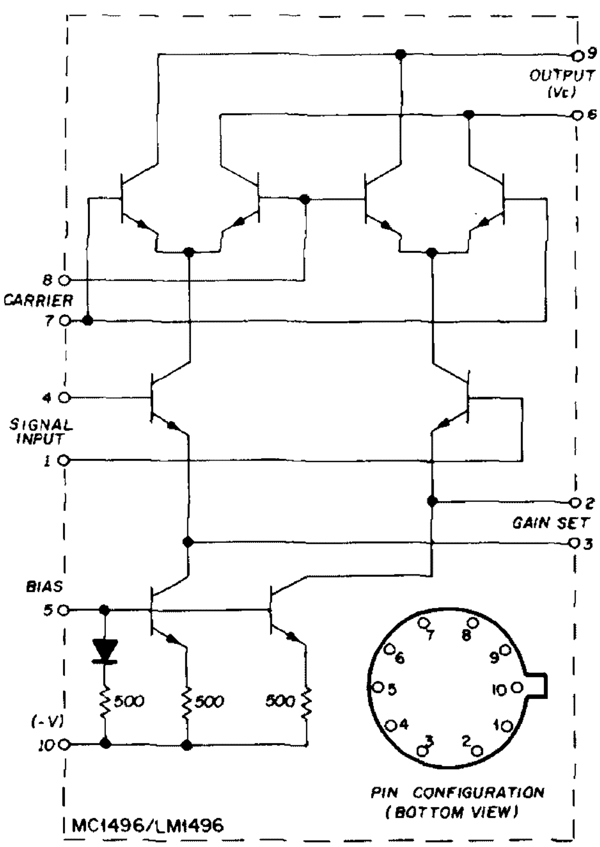


fig. 5. Schematic diagram of the MC1596/MC1496 balanced-modulator ICs.

for load mismatch so spare devices should be on hand.

About 1 watt PEP is available from the exciter without bypassing the emitters and that is a very conservative watt. However, even with an excellent signal and good signal processing it will still take perserverence and a good antenna to work stations. If your goal in life is QRPP, the mode best suited is CW, but a surprising number of stations can be worked with one watt of sideband.

The 1-watt PEP is more than adequate for driving the companion linear to a full 300 watts PEP input. As a matter of fact, a watt is too much drive for the linear and a 10-dB pad is used but the pad provides isolation that helps keep the exciter agc working properly. High-power linear transistors are still expensive but there is a flood of "illegal for class-D" linear amplifiers hitting the marketplace now and these linears are rated at more than 100-watts output. If these transistors work well at 27 MHz, they *will* work *nearly as well on all the* lower amateur bands and production quantities should be sizable judging from past reports on CB gear. In the meantime it might be well to consider a lower power linear or one with surplus vhf transistors.[7,8]

operation

The entire 1-watt transceiver draws a maximum of 800 mA on transmit and from 150 to 500 mA on receive. The power supply drain on receive can be reduced by switching the entire transmitter with the R/T switching transistor. The collectors of the 2N3866s need not be switched because they don't conduct with the bias removed. The main problem encountered in switching everything is that the bias on the balanced mixers takes about half a second to stabilize and a short burst of full carrier goes out over the air. This would be a small price to pay for portable operation and is not really noticable since the carrier is zero beat at the receiving station and would not normally be audible. However, vox operation is not recommended because the numerous times that the unit would cycle from receive to transmit would make the unit sound like slow CW with all the little bursts of carrier. For vox operation, only the bias to the last three transistors is switched. The audio and balanced mixers are left *live* and there is a signal being generated but it is isolated well enough that it is not heard.

It would be a shame to spoil an all

solid-state rig with an electro-mechanical monster for antenna switching. Full-fledged diode antenna switching would require standing off 80 volts or more of rf with the unit running at the 100-watt PEP output level. Where do you get 80 volts in a low voltage radio? There must be a simpler way — not a better way — just a simpler way to get around the use of complicated, noisy, unreliable hardware.

Since a lowpass filter was required anyway, it was decided to try to get some signal for the receiver from the lowpass filter. It seemed logical that one of the shunt capacitors in the lowpass filter could be lifted from ground and enough signal could be sampled to supply the receiver. Some tests were made and it was found that the middle capacitor of the filter was the best. In addition, it only requires standing off 3 volts of rf which is easily done with a single inexpensive diode. The voltage to switch the diode is fed down the coax from the receiver by inserting the voltage into the link on the receiver input tank coil and bypassing it to ground. This circuit will protect the receiver to some extent even if the dc control voltage fails. It also helps protect the receiver from strong local signals. This is a good system with multi-transmitter setups. The only bad part is the 10-dB of loss for the receiver, but the receiver is quite sensitive and signals on the lower bands are usually so strong that the rf gain is turned down to prevent cross-modulation anyway.

There may be a more elegant way of solid-state switching but this method requires only a resistor, a capacitor and a diode, about 15 cents worth of components. The 10-dB of loss could be greatly reduced if the filter were redesigned and the impedances were matched properly, but this is no trivial task and would best be done with a computer. Switching to transmit is swift and silent leaving the operator with the feeling that the rig just died, but after a few minutes on the air, confidence will be restored . . . if the 300-watt linear is being used.

The filter used to remove the harmonics output is one of the filters described in complete detail in a previous article.[9] These filters are probably more than adequate, but it is better to be overcautious. For the sake of simplicity, the same lowpass filter is repeated: once between the exciter and final power amplifier and again after the power amplifier. The filter between the exciter and final certainly could be of simpler design and still be adequate. Elliptic function filters are degraded drastically by severe mismatch and shouldn't be operated with vswr exceeding 3:1 at full rated power.

The dc voltages to the transmitter are switched by a medium power pnp transistor. As mentioned earlier, if low power consumption is desired, all power may be switched. If plenty of idle power drain can be tolerated, only the bias voltage to the last two stages need be switched. The collector voltages to these two stages may be left connected if desired. Positive voltage is sent through the primary of the receiver antenna coil, out through the receiver coax to the lowpass filter and the T/R switching diode. Positive voltage is also sent through a diode and 1000-ohm resistor to the receiver for muting.

The LM380 audio amplifier in the receiver can be muted by applying the voltage from the 1000-ohm resistor and diode to pin 2 but there will be some pop when the receiver is un-muted. If the receiver is improved by using the MC1590G (or MC1350P) to replace the Q4 stage, then the mute voltage from the resistor and diode is applied to pin 2 of the MC1590G (pin 5 if it's the MC1350P) and muting and un-muting is rapid and silent. This smooth and silent keying is a natural for vox operation. Some experimenting has been done with

"instant voice interruption" but very few amateurs have actually used it on the air.[10,11]

Since speaker operation with vox complicates things, there is no anti-vox included in the simple vox circuit shown in fig. 6. Operation with a boom-microphone type headset was the primary objective and it has worked quite well for that purpose. Surprisingly, it works fairly well with a speaker at moderate volume and using an Astatic 10-C microphone that is only slightly directional.

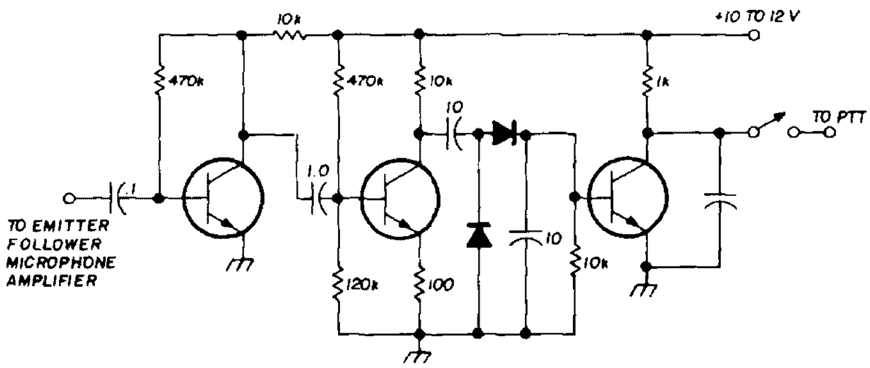


fig. 6. Vox circuit for use with headphones. Transistors are 2N3707, diodes are silicon such as 1N4154, 1N914 or similar.

It is desirable to apply a little of the receiver signal to the headphones when transmitting and the resistor in the mute control line can be increased to allow the amount of desired audio sidetone if the MC1590G or MC1350P is the controlled amplifier in the receiver. Audio sidetone is used in aircraft radios as an audible output indicator and to prevent the pilot from talking too loudly and overmodulating.

other frequencies

The transmitter is capable of operation at higher frequencies than the 10 MHz maximum limit arbitrarily imposed by the digital rf phase-shifter. The digital phase-shifter has provided the key to wideband operation and in itself will work well from the kilohertz region up to 10 MHz with nearly perfect phase quadrature. In fact, the Phase II transceiver may be the first ssb rig capable of operating on the experimental 160- to 190-kHz band. Of course, the ferrite transformers would have to be rewound with more turns (the same ratios) and a low-frequency vfo would be necessary.

At the other frequency extreme, it would be possible to use the rig at higher frequencies if phase-quadrature rf were derived from means other than digital ICs. The most promising and economical method is coaxial phasing lines similar to the phasing lines used for large antenna arrays. A quick check with a vector-voltmeter confirmed that above 14 MHz, complete phone-band segments can be covered with coaxial lines and the technique looks good for six and two meters. Coaxial lines are relatively insensitive to temperature variations that give other types of phase-shifters real problems. The coaxial line should be driven and terminated in the characteristic impedance of the line and the parallel paths of the two channels should be of the same impedance even if no coax is used in one side. Don't forget that the required quarter-wavelength line is an electrical quarter wavelength and must be multiplied by the velocity factor for the particular coaxial cable being used. For the twenty-meter band,

and to cover complete amateur bands, it would be necessary to include a small trimmer at the termination of the coax to trim from one end of the band to the other.

alignment

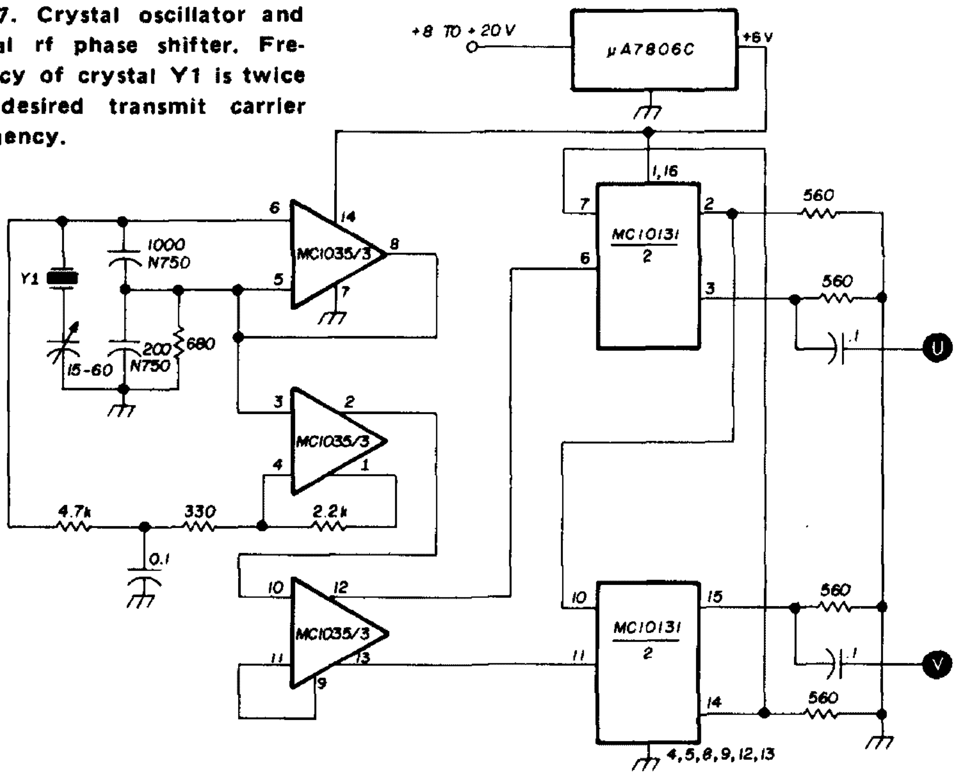
Alignment of the transmitter is very simple. Just adjust the two 50k pots for minimum carrier or minimum collector current then adjust the 2k pot on the balanced-modulator for maximum sideband suppression. It is best to use a separate receiver for both of these operations but it is possible to monitor the transmitters unwanted sideband in the receiver since the sideband selection is independent. The mute line must be disabled to monitor its own transmitter.

The carrier can be nulled by watching final collector current with no audio input or with the audio pot turned down and should be quickly checked after switching bands. Adjust the microphone gain pot (the 10k pot after the audio filter) so that some compression becomes obvious. This setting will have to be developed on an individual basis to suit each voice and amount of desired compression.

If further control of compression is desired the 15k resistor that feeds the agc signal back from the detector to the MC1590G can be altered as well as the value of the 56-μF capacitor at the detector. The 2N3866s should draw about 100 mA on voice peaks but it is possible to compress until the valleys are only about 80 mA but the audio sounds compressed at this level.

The audio sections, including the lowpass audio filter and the phasing networks, may be checked by applying the output from pin 1 of one balanced mixer to the horizontal input of an oscilloscope and the output from pin 1 of the other balanced-modulator to the vertical input of the scope. Both scope inputs should be sensitive enough to read one volt with good deflection. The

fig. 7. Crystal oscillator and digital rf phase shifter. Frequency of crystal Y1 is twice the desired transmit carrier frequency.



audio from the microphone will cause perfect, multiple, concentric circles. If the circles are slightly oblate but concentricity is good there is no problem because the gain of one balanced modulator is adjustable and will compensate for this gain error. If the circles are not concentric there is an error in the phase-shifter values or a misplaced component. If there are circles with no audio input, the unit is oscillating! Back up and try to isolate the *flying* stage. Oscillation is probably being caused by insufficient bypassing of the supply line or coupling through the power supply leads to the microphone preamplifier or the MC1590G.

If an audio signal generator is available it may be used to check the flatness of the filter response and the accuracy of the phase-shifter. One and only one circle will be present with a single audio tone fed into the microphone input. The circle should remain perfectly circular and of constant diameter with signal inputs from 300 to 2700 Hz.

performance

A great deal of time went into the breadboard just to clean up the audio and rf amplifiers and checking to make sure they were linear. Excellent signal quality has been the payoff and the ability to communicate better than the other guys who are running much higher power. A scope is a valuable tool for checking linearity, but both the rf and audio class-A amplifier stages have one simple check that can be made with a cheap voltmeter. The collector voltage should be exactly one-half the supply voltage.

An on-the-air roundtable with Collins, Heath, Swan, Drake and Phase II transceivers was recorded and the Phase II was definitely among the top two for quality. The unwanted sideband of the Phase II transceiver is garbled and unintelligible while the suppressed sideband of filter-type transmitters remains understandable even though attenuated. This leads people to believe that the sideband suppression of the Phase II is better than it actually is. It does, however, live up to the expectation of 40-dB unwanted sideband suppression.

The full transceiver with the 300-watt PEP input solid-state linear has been directly compared to a Heathkit Marauder, model HX-10, running comparable power. Input to the solid-state linear was held to 4 amps at 30 volts. Stations always rated the Phase II as "better copy" though most reported the same S-meter reading for both. Sideband suppression was equally as good as the crystal filter of the Marauder and some reports were in favor of the Phase II. When carefully questioned about the sideband suppression it usually amounts to "unintelligible."

Direct conversion is the ideal experimental transmitter for the amateur because there are fewer and more predictable spurious responses from direct conversion. The only spurs likely are harmonics and they are very effectively removed by the lowpass filtering. Conventional superheterodyne-type transceivers have possible images at the intermediate frequency or frequencies, depending upon the number of conversions. Of course, there is always a chance of parasitic oscillations in either the audio or rf of a direct-conversion transceiver, but these are more easily recognized and easier to troubleshoot with the direct-conversion method.

The thing that saves a lot of troubleshooting time is that the same spurs will appear in both the receiver and transmitter. For example, if the vfo being used has a parasitic and is effectively supplying two separate and nonharmonically related signals to the rf digital phase-shifter, two separate signals will be received and two separate signals will also be transmitted. However, don't get too excited about receiving two signals at once. Not only the two signals

but the sums, differences, and an endless collection of products also enter in and make it impractical.

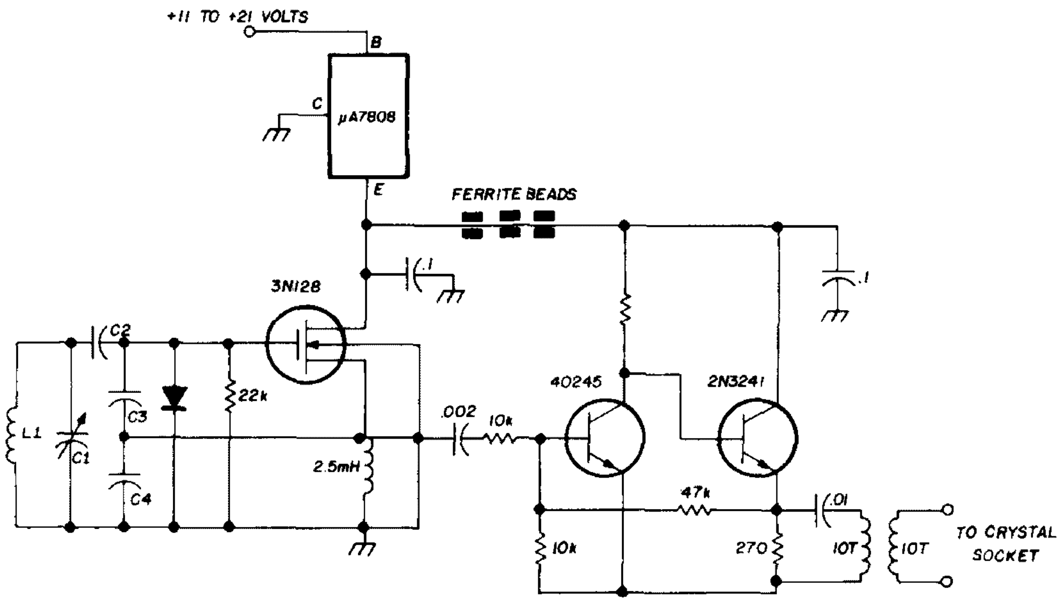
The transceiver always listens where it talks, providing the receiver input is broadband like the transmitter, but it is usually better to have a little more selectivity in the receiver to protect from out-of-band signals. It is also very important to have a clean vfo signal.

variable-frequency oscillator

Close-in noise in the vfo signal is more important in this case than the harmonics that might be present. The close-in noise on a vfo may be caused by fm of the oscillator and that in turn is caused by poor voltage regulation. Close-in noise is a particular problem when using a frequency synthesizer with a phase-locked, voltage-controlled-oscillator. The output signal tends to be fm at the reference oscillator frequency and this should be carefully investigated by those intending to use a synthesis scheme.

For those of you wishing to use the Phase II reciter as a transmitter paired with some other receiver, the rf digital phase shifter and crystal oscillator schematic is shown in **fig. 7**. The same scheme can be used with other ICs, but a dual-D flip-flop should be used to help preserve good tracking with temperature changes.

A good vfo is necessary for amateur-band use except in the rare cases when a crystal is used for net operations. The vfo shown in **fig. 8** was originally described in *QST* and has appeared in many transmitters and transceivers over the last eight years.[1,2] The mosfet is an



operating frequency	vfo frequency	L1*	C1	C2	C3	C4
1.8-3.0 MHz	3.6-6.0 MHz	30 turns	250 pF	470 pF	820 pF	820 pF
3.0-5.0 MHz	6.0- 10 MHz	25 turns	200 pF	390 pF	680 pF	680 pF
5.0-8.8 MHz	10 - 16 MHz	14 turns	75 pF	220 pF	470 pF	470 pF

*L1 wound on Amidon T-68-2 toroidal core; use as large wire as possible for full, single layer.

fig. 8. Variable-frequency oscillator for the Phase II reciter. Values for L1 are approximate and are determined by the exact value used at C1. Value for C1 is approximate with total made up with fixed capacitors to cover desired band segment. Toroidal core for transformer T1 is Ferronics 12-360J (or Amidon T-50-3).

extremely stable element and the addition of a toroidal inductor makes it more compact. An MC7808 or μA7808 IC voltage regulator provides stable 8-volt regulation and filtering which is important for the elimination of fm and drifting in any vfo.

The most important part of the vfo and the most difficult to locate is the variable capacitor. Good variable capacitors just are not used much in commercial equipment anymore and the really good ones invariably are salvaged from some old piece of tube-type military gear; they have plate spacings of 1/16 inch (1.5 mm) or more and are mechanical monsters. That is the price of progress. Some mechanical genius should start marketing a log-variable inductance similar to the units used in the famous Collins linear-permeability-tuned-oscillator (PTO) and solve all our problems.

Almost any value variable capacitor can be adapted for vfo use by using series and parallel silver-mica capacitors to set the range and frequency. The values given in **fig. 8** are only ball-park figures. The L/C ratio of this particular vfo can be varied quite a lot without degrading the performance. An hour or two with a slide-rule, calculator, or ouji board will allow the use of practically any capacitor that comes out of your junk box.

Also consider the use of a high-frequency vfo at 40 MHz or higher and dividing down to the desired frequency with digital ICs. The stability of the higher frequency vfo is notoriously poor but it is as good percentage-wise as the lower frequency counterpart and the frequency errors will be divided along with the frequency. The higher frequency vfo will allow use of smaller components and more readily available variable capacitors.

output filtering

A Drake TV100-LP or TV1000-LP lowpass filter should follow *any* transmitter, be it commercial or homebrew and particularly a transmitter with a broadband linear. The Drake filter does an exceptional job of suppressing TVI, is recognized by the FCC, and cannot be duplicated in the average hamshack. The lowpass filters used in the Phase II reciter are not designed to be TVI filters as such because the silver-mica capacitors used will display a resonance somewhere in the vhf region and filter performance will deteriorate rapidly. If better vhf filtering is desired from the lowpass filter, the capacitors going to ground should be made up of two or more smaller capacitors of differing values and even different types of capacitors, such as ceramic.

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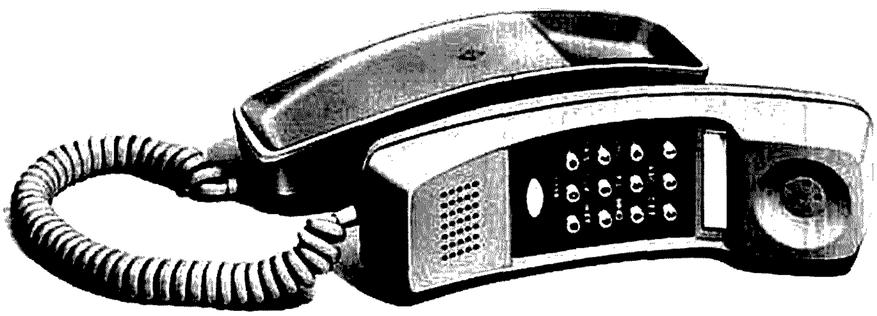


converting the slim-line touch-tone handset

A telephone handset combining touch pad, microphone and speaker into a single unit

Joseph M. Hood, K2YAH, Rochester, New York

Tone signalling for amateur repeater autopatch access, repeater secondary control access, or selective calling is a rapidly expanding technique in fm circles. Many amateurs are using a touch-pad mounted in a separate enclosure as a source of these tones. While this method is acceptable, it does have disadvantages; finding a place for the bulky touch-pad enclosure and switching the associated microphone and push-to-talk circuit interfaces with the transceiver to make the touch-pad approach somewhat inconvenient.

A solution to this inconvenience is to replace the separate touch-pad and microphone elements with a unit which contains both, a *Slimline*, Touch-Tone*

*Touch-Tone is a Registered Trademark of the Bell System.

handset. The handset also has the additional advantage of an earphone, which can be used for private listening when operating with a car full of sleeping children or other persons you don't wish to annoy with the receiver audio.

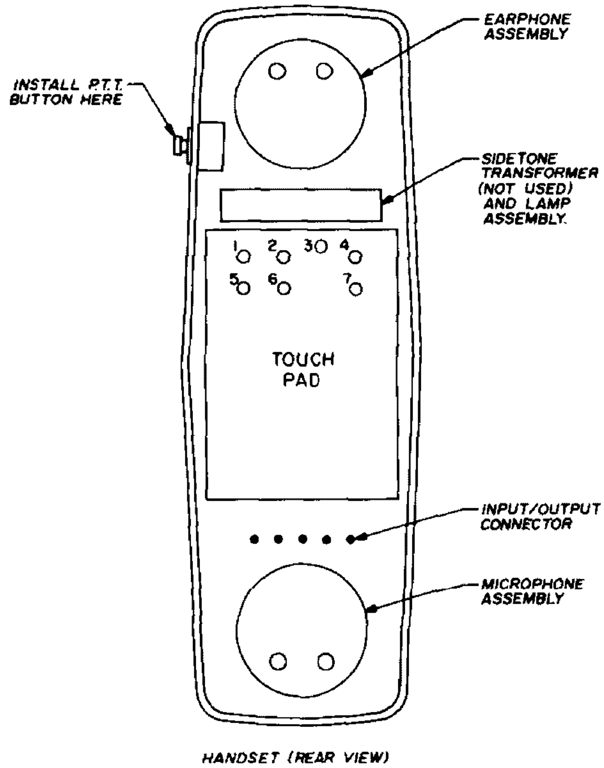


fig. 1. Rear view of the handset showing suggested location for mounting the PTT switch.

modification

The *Slimline* phone is not very useful for amateur radio use as wired for telephone service and will require some electrical and mechanical surgery to make it compatible with fm transceiver input/output circuitry. The modification is begun by opening up the unit by removing the two screws found under the removable, transparent plastic, telephone number cover. With the unit open, you can see the plastic printed circuit which contains the interwiring for the handset. This printed circuit must be removed by extracting the screws on the earpiece, touch pad, and microphone connections, and then carefully unsoldering the remaining solder connections. Don't lose the screws as they will be used when the unit is rewired.

Since the handset has no push-to-talk button, one will have to be installed. The first thing you will notice is that there isn't much room for one anywhere in the unit. However, a miniature pushbutton switch, available from Lafayette Radio and Electronics (part number 99P62184), will fit in the space between the earphone assembly cover and the side of the handset, as shown in **fig. 1.** Care should be taken in locating the specific spot where the switch will be mounted. When this is determined, drill a small guide hole in the handset case using a slow drill speed. Then follow with a drill large enough to allow switch mounting. Again, use a slow drill speed and exercise care in drilling. When the switch is installed the handset is ready for rewiring.

rewiring

To use the handset it must be wired as shown in **fig. 2.** This circuit uses contacts inside the touch-pad to switch the microphone element in the audio

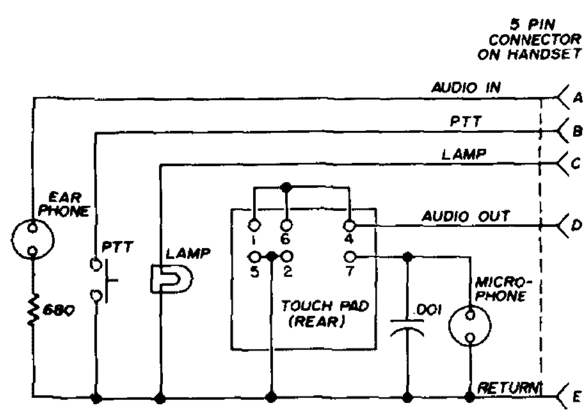


fig. 2. Internal rewiring of handset. Use an ohmmeter to check input/output connector pins for continuity.

output line when no touch-pad buttons are depressed; conversely, the touch-pad's output replaces the microphone's whenever a touch-pad button is actuated. The output connector on the telephone is a tricky area. When you view its connection points and output pins there seems to be a one-to-one, geometric association. Not true! Some of the pins do *not* connect with their most adjacent input point; use an ohmmeter to check which pin goes to which output point and you won't be fooled.

interconnecting

The rewired handset may be connected directly to any transceiver having an audio input compatible with a carbon microphone. This connection is shown in **fig. 3A**, and may also be used with transceivers designed to use dynamic microphones having integral preamplifiers. If your radio uses a low impedance (500 ohm) dynamic microphone without an integral preamp, the circuit of **fig. 3B** should work. If the rig uses a high-impedance ceramic microphone the circuit of **fig. 3C** is suggested. The potentiometer should be adjusted to give the same modulation level as the original microphone.

The input/output to the radio may be made with the coiled telephone handset cable. However, you will find this cable is very difficult to solder. The conductors are a combination of coiled copper and cotton thread; this makes them very flexible but makes soldering to them somewhat tedious. However, by carefully removing the thread and keeping the heat to a minimum, it can be done.

The connector used at the transceiver input should be one which provides effective strain relief for the handset cable. If a good strain relief isn't provided, the small coiled copper wires, which become brittle when soldered, will surely break. An Amphenol type 91-MC6M cable plug and 91-PC6F chassis receptacle are recommended.

The handset approach to fm mobile operation relieves the "where to put the

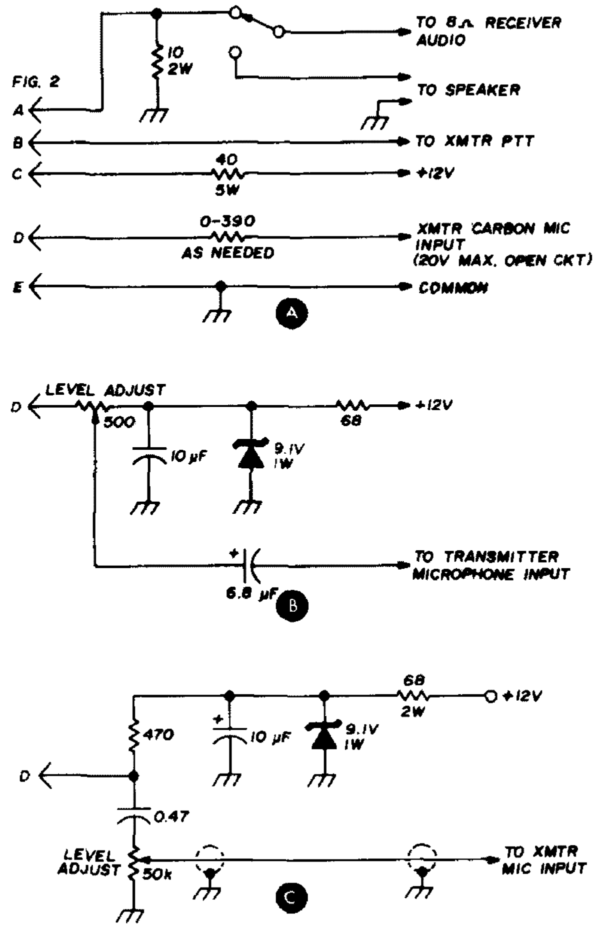
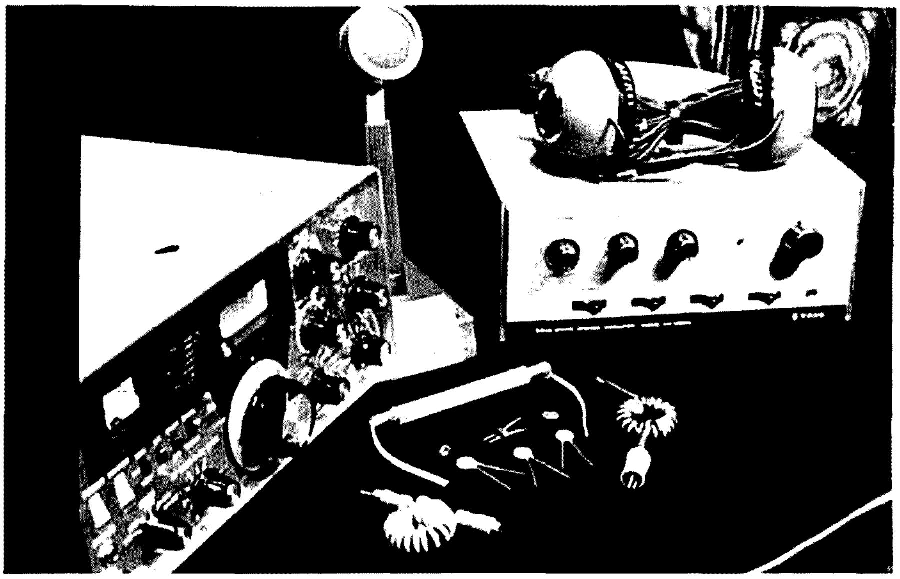


fig. 3. Interconnection wiring between handset and transceiver. For carbon microphone or amplified dynamic microphone inputs use the circuit of (A). Dynamic microphone inputs (500 ohm) require the circuit of (B). Use (C) for inputs designed for ceramic microphones. Other wiring is the same for all four types.

touch-pad" dilemma in first-class style. The carbon microphone element in the telephone produces surprisingly good audio quality, and the private listening feature is a nice bit of serendipity which is quite useful. If nothing else, the handset will enhance your "Frank Cannon" image immeasurably.

ham radio



audio-radio frequency interference — its cause and cure

Most cases of rf interference to audio hi-fi and stereo equipment can be cured — typical solutions are discussed here

Harry Leeming, G3LLL*

Having an active hobbyist's and a commercial interest in the fields of amateur radio as well as in stereo and high-fidelity equipment, I am very concerned about hi-fi interference and its possible cures. It would seem that many audio equipment manufacturers leave their products wide open to the reception of unwanted radio signals, and to determine to what extent, I decided to run some tests.

An amateur single-sideband transmitter with about 200 watts power output was connected to a center-fed dipole with a coaxial transmission line; the antenna and transmitter were located about 50 feet (15 meters) from my company's audio equipment showroom. Test transmissions were made at both 21.2 and 3.6 MHz. With no external inputs (phono pickup or tuner), half of the fourteen amplifiers tested proved susceptible to interference on 80 meters, and five suffered interference on 15. When the phono pickup

*Holdings Photo Audio Centre, Mincing Lane, Darwen Street, Blackburn BB2 2AF.

was connected the interference became quite severe on nine amplifiers, and when the tuner and antenna were connected, all amplifiers except one suffered heavy interference.

Since the tests were made in our hi-fi showroom the speaker leads were rather long, and possibly resonant near the 80-meter band, so this probably explains the greater incidence of interference on the lower frequency. Shorter leads would possibly have reduced interference on 80 meters and increased it on 15.

I think it would be fair to say that the circumstances and power used were pretty average, and it would be reasonable to expect similar results within a few doors of an amateur radio station, or within less than a mile or so of a high-powered commercial station. In all of these tests, and in some more severe tests undertaken at home, the amplifier which came out best was the Quad which, from an examination of the circuit diagram, was found to have considerable built-in RFI protection as well as being completely enclosed in a metal cabinet.

In the future, as radio transmitters for broadcasting, business radio and amateur radio multiply in power and number, and the RFI rejection of amplifiers gets worse (due to the advent of the transistor and the printed circuit), the problem needs immediate attention. Fortunately a few manufacturers are now

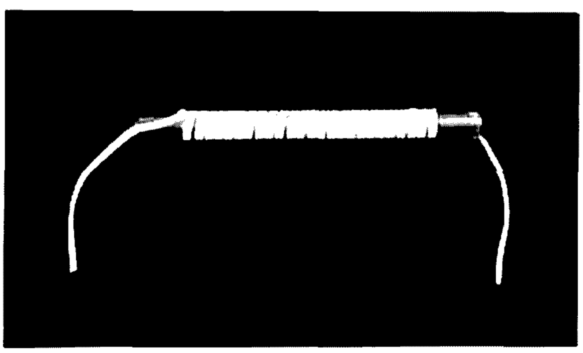


fig. 1. Loudspeaker lead can be formed into an rf choke by winding the lead around a ferrite antenna rod.

taking note of the problem; it is to be hoped that others will soon follow suit.

external pickup

Before delving into the equipment itself, let's see what can be done externally to help when unwanted radio transmissions are already being picked up. The first move is to determine the frequency of the transmitting station, as knowledge of this will enable you to make a more intelligent approach to the problem. Calculate the wavelength of the interfering signal and take a look at the lengths of the various leads used in the audio installation. A lead only a fraction of a wavelength long (say 1/20th) makes quite a good antenna, and any lead which is a quarter-wavelength long (or any multiple thereof) will make a very effective antenna.

If no work is contemplated inside the amplifier, the only possible approach is to try to establish how the RFI is entering the circuit. A simple step is to remove the input leads one at a time, and note which one reduces or eliminates the interference. The speaker leads are quite likely causes of trouble and hence, experimentally, they can be shortened to less than 1/20th of a wavelength, or can be disconnected in favor of a pair of headphones with short leads.

If the RFI is still present with the leads all disconnected, the trouble is due either to rf pick-up on the amplifier's internal wiring, or it is coming in through the power lines. Check for interference arriving via the ground lead by disconnecting it. To test for pick-up from the power lines simply pull the plug out quickly while the interference is manifesting itself. If RFI is entering by the ac line it will disappear the instant the plug leaves the socket; if it is being picked up in the internal wiring of the amplifier it will slowly fade away as the power-supply filter capacitor discharges.

If the interference is arriving through

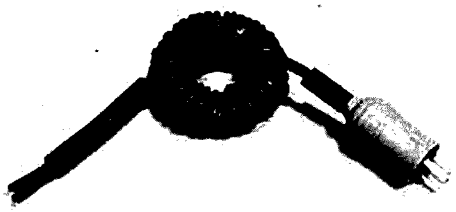


fig. 2. Ferrite or powdered-iron toroid can also be used to form in-line rf chokes with loudspeaker leads.

the ac power lines, install an ac line filter which is effective at the frequency of the unwanted transmission. This type of filter is used to suppress vacuum cleaners, electric shavers, etc., and is available from most electronic distributors. Alternatively, the hot, neutral and ground leads can be treated with ferrite cores, as is described later in connection with speaker leads.

speaker leads

If RFI via the speaker leads is the trouble, check that the leads are not a multiple of a quarter wavelength, and if they are, alter the length, preferably by shortening. If this does not cure the trouble, connect four 0.01-μF disc ceramic capacitors from plus and minus speaker sockets, with the shortest possible connections, directly to the chassis. If the interfering signal is lower than about 5 MHz, larger capacitor values may be required, but they should be used with caution in case they affect the high-frequency stability of the amplifier.

These moves should almost certainly reduce the strength of interference, but if it is still present, form the speaker leads near the amplifier into radio frequency chokes (fig. 1 and 2). This can be done by wrapping the lead around a ferrite antenna rod, or better still, by winding the lead around a ferrite core. About twenty turns will be needed to form an effective choke for frequencies in the range of 10 to 20 MHz, with proportionally more turns for lower frequencies.

Reception on the pickup leads themselves is only common at higher frequencies where the leads begin to form an appreciable fraction of a wavelength. If possible, the simplest solution is to shorten the leads. Alternatively, the pickup leads can be wound around a ferrite rod or ferrite core to form an rf choke (fig. 3). While investigating this side of the problem, check the pickup grounding as the wiring scheme shown in fig. 4 sends any signal picked up on the ground lead straight into the phono socket of the amplifier. Fig. 5 is a much better arrangement.

tuners

If RFI via the tuner leads appears to be the trouble, try disconnecting the fm antenna as the coaxial feedline makes an excellent shortwave antenna with the signal going to ground through the tuner leads and the amplifier printed-circuit boards. The answer here, as is shown in fig. 6, is to provide a 1:1 transformer in the antenna lead which will pass the vhf fm signal but isolate the lower-frequency interfering signal. Fig. 7 shows a simple way of doing this with a slight loss of signal; if you cannot afford to lose a little signal the transformer

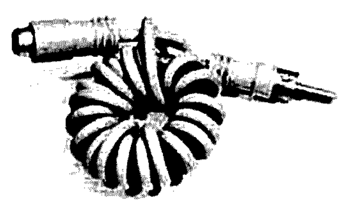


fig. 3. In some cases interference can be reduced by winding external audio input leads on toroidal cores.

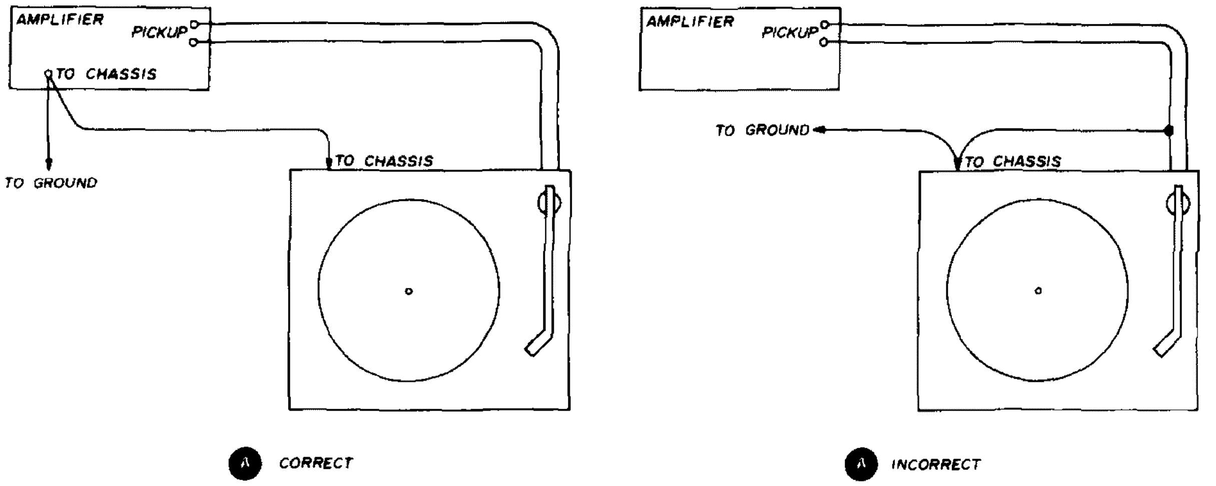


fig. 4. Correct grounding connection for external pickup is shown in (A), improper ground connection is shown in (B). In the circuit of (A) the only input signal is the desired one, while in (B) any signal picked up on the ground lead goes directly to the input socket of the amplifier.

shown in **fig. 8** can be made quite simply.

If the feedline is not grounded, a 1-megohm static discharge resistor should be connected between the primary and secondary of the transformer. If you are not able to purchase the small ferrite core, it can be obtained by dismantling a balun transformer from an old TV set.

If the interference continues, even with the fm antenna disconnected, the audio connecting leads can be treated with a ferrite core as described previously.

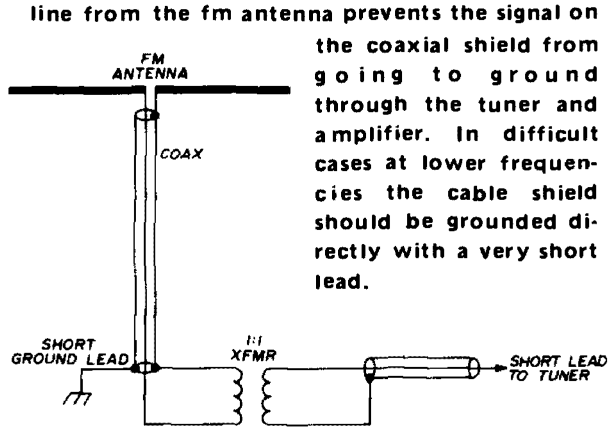
equipment modifications

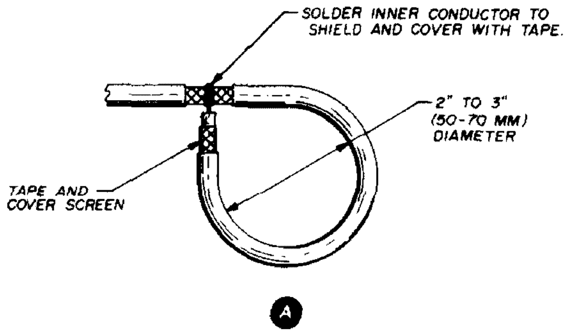
In addition to the external connections discussed so far, I decided to see what internal modifications would be needed to improve an amplifier's performance from the RFI point of view. I quickly realized that more RFI was getting in through the shields of the input leads and through the ground side of the speaker leads than was traveling in via the live conductors. To reduce problems with hum caused by ground loops, most good quality amplifiers do not have the speaker or input connectors grounded near their mounting points, but return them to the chassis through the circuit-board panels. This arrangement is fine from an audio point of view, but does nothing to stop rf from entering the circuit.

The problem was investigated further with the use of an rf signal generator and it was soon realized that no single component was going to cure the trouble, and a "belt and suspenders" attack was decided on.

Rf interference to audio amplifiers is caused primarily by the transistor junctions rectifying, and therefore demodulating, the radio frequency energy. The answer here is to short-circuit the junction sensitive to radio frequency by the use of bypass capacitors. All the transistors in my preamplifier seemed sensi-

fig. 5. The 1:1 transformer in the transmission line from the fm antenna prevents the signal on the coaxial shield from going to ground through the tuner and amplifier. In difficult cases at lower frequencies the cable shield should be grounded directly with a very short lead.





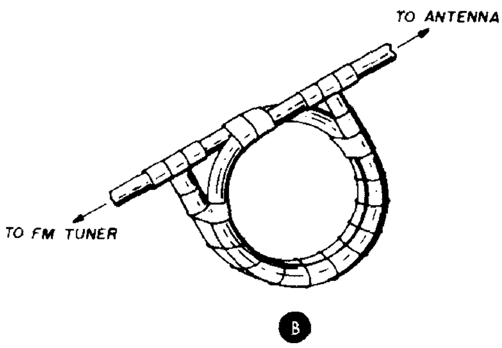


fig. 6. Details of Faraday double-loop filter are shown in (A). Two of these loops are placed next to one another as shown in (B), taking care to insulate all wires and shields. Be sure the two loops are laced or taped firmly together.

tive, and in the end it was decided that they should all be bypassed. As has been noted previously, much of the rf was entering through the shields on the phono and tuner leads, so these were also bypassed to the chassis; it was not found necessary to bypass the speaker leads once the preamplifier had been attended to.

The results of a handful of bypass capacitors were so successful that it is now possible to operate the amplifier with only very slight breakthrough with a high-power transmitter and antenna in the same room as the amplifier. Previously, with the two units 100 feet (30 meters) apart, the interference was at a deafening level.

Successful as the modification seemed from tests in my company's workshop, it was decided in the interests of science (and amateur radio) to see what happened when the equipment and transmitter were operated in my own home. When the modified amplifier was first connected, results with the transmitting antenna 20 feet (6 meters) away were rather disappointing as interference still varied between "quite loud" and "deafening," depending upon the operating frequency.

The interference was not too bad when the tuner was switched in, but was very objectionable when switched to the phono pickup. Further investigations showed that the trouble disappeared if the fm antenna (which was only about 3 feet [1 meter] from the transmitter antenna) was disconnected. The feedline to the fm antenna was fitted with a 1:1 transformer, and while I do not normally make a habit of listening to my hi-fi setup when talking to someone on the other side of the world, this would now theoretically be possible! Of course, the signal injected into the amplifier at this range is more than you would normally expect in the average home, but it does illustrate that many manufacturers could considerably improve their equipment in this respect at negligible cost.

bypass capacitors

Those readers contemplating similar modifications should remember that there is no such thing as a perfect capacitor; all have some inductance. In general, the larger the value of the capacitor, the larger its self-inductance, so the theoretical circuit of a practical

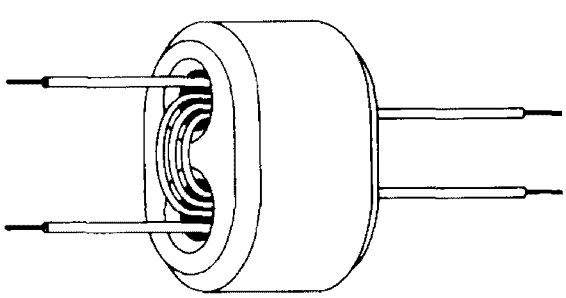


fig. 7. Simple 1:1 transformer uses two-hole ferrite bead salvaged from old television set.

capacitor appears as a series LC circuit as shown in **fig. 9**. At high radio frequencies, the inductance of, say, a metal-foil capacitor can have a reactance of hundreds or even thousands of ohms, making the capacitor useless when it comes to bypassing such radio frequencies.

If you are to effectively short out the base-emitter junctions of transistors, or bypass the shields of connecting cables to the chassis, the reactance of the capacitor you use should not be more than an ohm or two at the frequency in question. A very suitable capacitor for

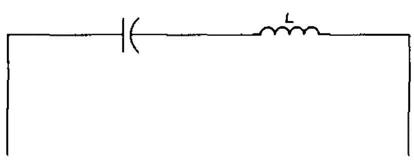


fig. 8. A practical capacitor has unavoidable inductance of leads as well as self inductance. At high frequencies inductance must be kept to an absolute minimum.

this type of work is the disc ceramic capacitor, which has a very low self-inductance. From an RFI point of view a 0.01 μF disc ceramic capacitor with minimum-length leads will present a reactance of only a few ohms from about 5 to 100 MHz.

At lower frequencies higher value capacitors may be needed, but they may affect response at the highest audio frequencies so unless the interference is severe from a transmitter on 160 or 80 meters, capacitors of lower value should be fitted between the base and emitter junctions. In practice, values between 500 and 5000 pF appear in the circuits of manufacturers who have taken precautions in this direction.

If the trouble is with breakthrough of television audio or business radio transmissions in the 50-200 MHz range, a smaller value capacitor, say 50-200 pF, may be even better; when connected with the minimum practical lead length it is possible for such a capacitor to form a series-resonant circuit to provide an almost dead short.

Whether or not a particular value capacitor will affect the audio response of an amplifier is difficult to predict, as much depends upon the impedance of the circuit. When adding components it is advisable first to modify only one channel of a stereo amplifier, so that the square-wave and frequency response can then be compared with the unmodified channel to ensure that the aduio response has not been upset.

At frequencies beyond 100 MHz or so, it becomes increasingly impractical to add capacitors to a circuit with short enough leads for them to be really effective. At this frequency the best approach is to fit ferrite beads on to the transistor leads. These beads increase the inductance of the transistor lead and operate as an rf choke.

summary

Since doing the tests and modifications to my own amplifier, several cases have occurred where customers have had serious trouble with rf interference, and the following modifications have always produced a cure:

1. Input connectors. Install a 0.01 μF disc ceramic capacitor from ground side of all input connectors to chassis.

2. Transistors. Install a 1000 pF disc ceramic capacitor from base to emitter of all transistors in the preamplifier.

3. Loudspeaker terminals. Install a 0.01 μF disc ceramic capacitor from the live and ground side of these terminals direct to the chassis.

While modification **3** will be found quite effective on its own, as it can often be done externally, it has not generally been found necessary when the other work is done internally.

ham radio

500-MHz decade prescaler

Using new sub-nanosecond ICs to build a ten-to-one prescaler that will extend the frequency range of your counter to 560 MHz

Wayne C. Ryder, W6URH, 115 Hedge Road, Menlo Park, California 94025

About seven years ago, using the new emitter-coupled logic (ECL), I built a 100:1 digital prescaler that extended the range of my 1-MHz vacuum-tube counter to 100 MHz. Although the prescaler stopped well short of two meters, it was a significant improvement at the time. Recently I removed the old 100:1 prescaler circuitry from the chassis and installed a decade prescaler circuit that accurately counts to beyond 500 MHz. All that was required was two ICs and about two hours of bench time.

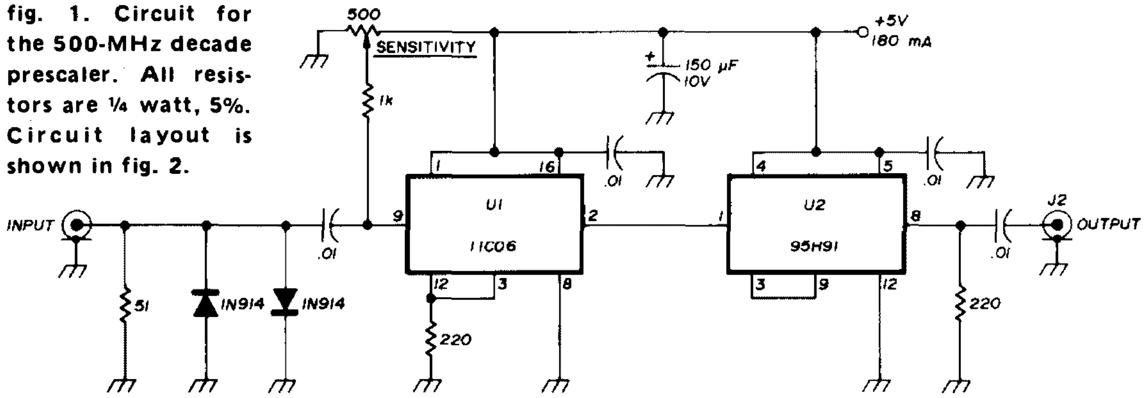
circuit

The circuit for the 10:1 500-MHz frequency scaler is shown in **fig. 1**. The first IC, a Fairchild 11C06, is a type-D flip-flop rated to 700 MHz by the manufacturer.[1] However, with the circuit layout I used, the prescaler stops counting properly at about 560 MHz. The 95H91 is a Fairchild divide-by-5 counter IC from the same ECL family as the popular 95H90.* Input sensitivity of the prescaler is less than 100 mV from 10 to 500 MHz.

The two back-to-back diodes from pin 9 of the 11C06 to ground protect the input against overload. Nevertheless, the maximum input voltage should not exceed 1 volt rms. The 500-ohm pot sets the bias voltage on U1 to about 3 volts. This control should be adjusted for maximum sensitivity. The output from pin 8 of the

*Total price of the two ICs is about $36 in small quantities from your local franchised Fairchild distributor. Motorola and Plessey manufacture similar sub-nanosecond ICs, including several that are rated to 1000 MHz.

fig. 1. Circuit for the 500-MHz decade prescaler. All resistors are ¼ watt, 5%. Circuit layout is shown in fig. 2.



95H91 is connected to a 50-MHz frequency counter. If the connecting cable is more than about 12 inches (30cm) long, a 50-ohm termination should be used.

construction

In my prescaler all the components are installed on a small piece of copper-clad circuit board about 2-inches square (50x50mm). I used point-to-point wiring as shown in the layout in fig. 2. Make sure all the component leads are as short as possible.

In my unit I included a regulated +5 volt power supply for convenience (see fig. 3). Although the prescaler requires 180 mA and the Triad T90X transformer is only rated at 100 mA, it remains quite cool, even during extended periods of operation. If you are going to use this prescaler with a TTL-based counter that has a regulated +5 volt supply with sufficient reserve, you may be able to power the 500-MHz prescaler from the existing supply. Make sure, however, that your counter's +5 volt supply can handle the additional 180 mA load.

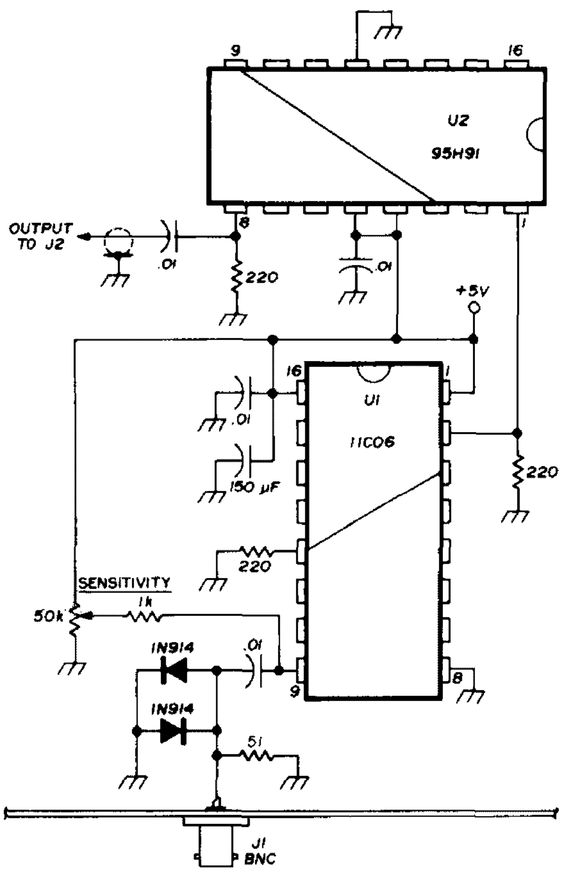


fig. 2. Component layout for the 500-MHz decade prescaler. Circuit is built on small section of copper-clad circuit board. All leads to U1 must be as short as possible.

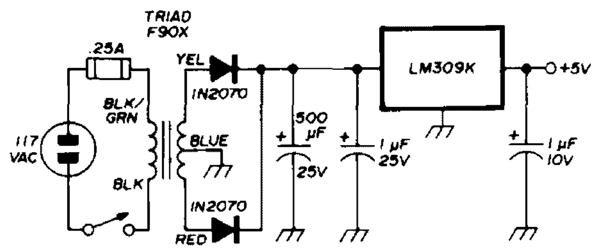


fig. 3. Regulated +5 volt power supply for the 500-MHz prescaler.

reference

1. Doug Schmieskors, WB9KEY, "1200-MHz Frequency Scalers," *ham radio*, February, 1975, page 38.

ham radio

stable crystal oscillators

A selection of stable crystal-oscillator circuits for 50 kHz to 80 MHz which minimize the influence of the transistor and the crystal's series resistance

Ulrich Rohde, DJ2LR, 14 Gloria Lane, Fairfield, New Jersey

Crystal oscillators are commonly used wherever stable frequencies are required. The crystals selected for these circuits are not always completely known so far as their inner parameters are concerned, and many radio amateurs run into difficulties with the crystal's equivalent series resistance, R_s (if it is too high, many published oscillator circuits do not operate properly).

Since there is no practical oscillator circuit which will cover the entire frequency range from, say, 50 kHz to 150 MHz, it is the purpose of this article to describe some circuits, using transistors, which are fairly independent of crystal losses and will give extremely stable frequencies. Almost any general-purpose rf transistor may be used in these circuits provided that the transit frequency, f_T, is higher than 250 MHz.

500 to 800 kHz

Fig. 1 shows an oscillator circuit which can be conveniently used in the range from 50 to 800 kHz. The only requirement for the crystal is that it must be operated in its fundamental series mode. The adjustable trimmer capacitor permits enough pulling range, and the 0.7-volt rms output is more than adequate for most requirements. In most cases the frequency range of this oscillator can be easily extended to 100 kHz, using crystals widely found in calibrator circuits.

1 to 20 MHz

Fig. 2 shows an oscillator circuit which exhibits extremely low power dissipation in the crystal, thus giving ultimate frequency stability. Capacitors C1 and C2 must be selected according to the frequency range as shown. While most designers use almost the same value capacitor at both C1 and C2, the capacitance of C1 should be substantially higher than C2. This reduces the influence of the transistor on the stability of the circuit by more than five times. For some unknown reason, only a few people are apparently aware of this advantage.

In the circuit of fig. 2 the output voltage is taken across the parallel RC circuit made up by capacitor C3 and

resistor R1, a 22-ohm resistor. Together with the crystal R1 and C3 form a low-pass filter which suppresses the second harmonic by 60 dB.

In cases where high stability must be combined with the selection of many channels, the oscillator circuit and diode switching scheme shown in **fig. 3** is highly recommended. In this circuit the crystals are used in their series-resonant mode, and depending upon the parallel capacitance of the fixed trimmer capacitor, 39 pF in the schematic, the individual frequencies may differ substantially from crystal to crystal. In this circuit, as well as in the others presented here, the influence of the external components is minimized.

harmonic oscillators

Harmonic oscillators are used for higher frequencies. It is very difficult to build stable crystal oscillators using 5th or 7th overtone crystals because these crystals are usually BT cuts which result in very poor temperature coefficient characteristics. It is much more convenient to use an AT-cut, third-overtone crystal and take advantage of the inherent frequency-doubling capabilities of the transistor. These harmonic oscillator circuits are often found in mobile radio systems where a number of channels are required.

The circuit in **fig. 4** shows a switch-

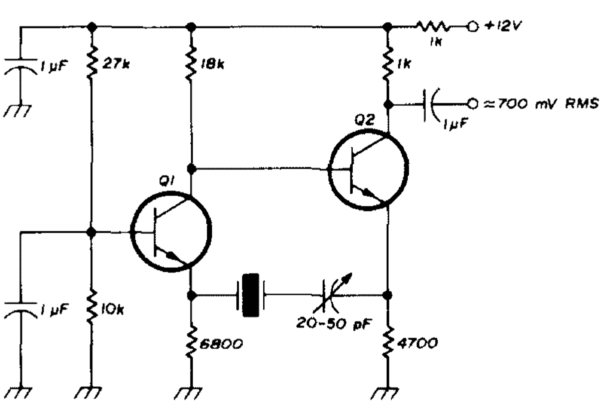
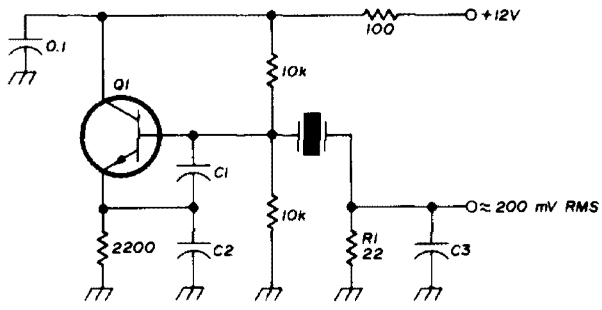


fig. 1. Crystal oscillator circuit for use over the frequency range from 50 to 800 kHz. Transistors Q1 and Q2 are types 2N708, HEP50, BC108 or similar.



FREQUENCY	C1, C3	C2
800 kHz 4 MHz	2200 pF	560 pF
4 MHz 20 MHz	390 pF	100 pF

fig. 2. Crystal oscillator circuit for the frequency range from 800 kHz to 20 MHz uses series-resonant. Fundamental-mode crystals. Parallel circuit made up of R1 and C3 suppresses second harmonic by 60 dB. Transistor Q1 is a 2N708, HEP50, BC108 or similar.

able overtone oscillator. The crystals, third-overtone types, may oscillate between 20 and 80 MHz. The series inductance, not required for each crystal, must be selected so that it is series resonant with 10 pF at the crystal's operating frequency. The total number of switchable channels may be as high as twenty, and the circuit will still remain stable without showing any uncontrollable oscillations.

The tuned circuit in the output of **fig. 4**, which can easily be modified into a bandpass filter from the single-tuned circuit, will provide about 500 mV rms into 50 ohms at two times the crystal frequency. If a bandpass filter is used at the output, subharmonic suppression will be greater than 60 dB.

Fig. 5 shows an overtone crystal oscillator circuit which can be either modulated or used as a very stable vcxo. In circuits where the final frequencies are derived by mixing one oscillator output with another, frequency adjustments can be made externally by applying a dc voltage.

For example, assume a single-conversion receiver is to be built for the 144-MHz band which will work only at

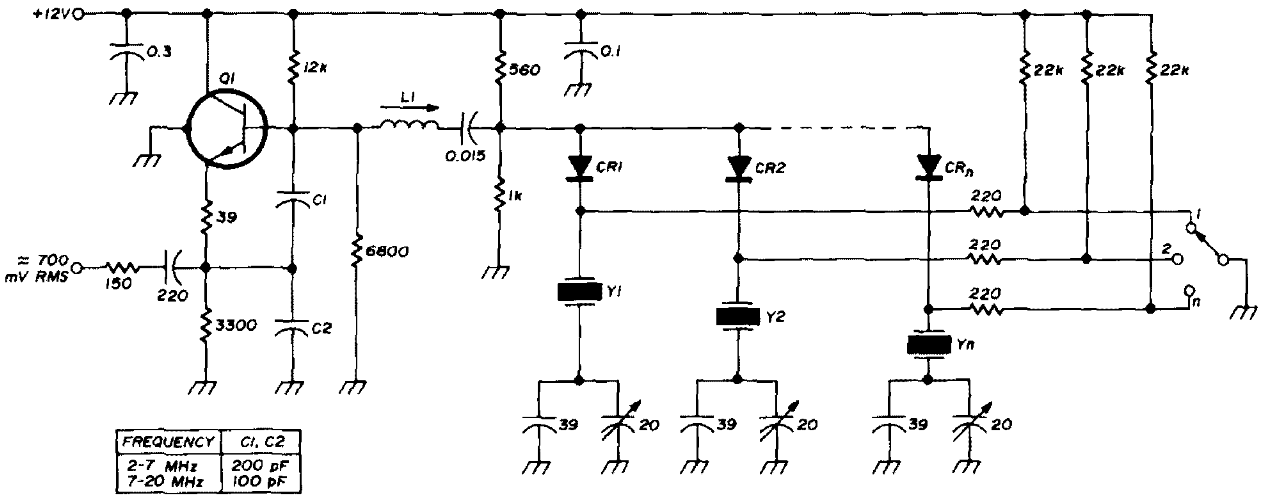


fig. 3. Switchable crystal-oscillator circuit for use over the frequency range from 2 to 20 MHz uses series-resonant, fundamental-mode crystals. Inductance at L1 should be about 30 μH at 2 MHz, and about 1 μH at 20 MHz. Transistor Q1 is a 2N708, HEP50, BC108 or similar rf npn type. Diodes CR1, CR2 thru CRn are switching types such as the BAY67.

the ssb portion. Most ssb stations operate near 144.1 MHz, so if a 58-MHz oscillator is used and doubled in frequency to 116 MHz (for use with a 28-MHz i-f) the electronic tuning (pulling) range is about 60 kHz, more than enough range for typical two-meter ssb operation. There is practically no noticeable sacrifice in frequency stability if the tunable frequency range is not extended beyond 60 kHz.

Similar oscillators may be useful in portable high-frequency transceivers where the pulling range may be slightly less but still sufficient to cover the CW portion of an amateur band.

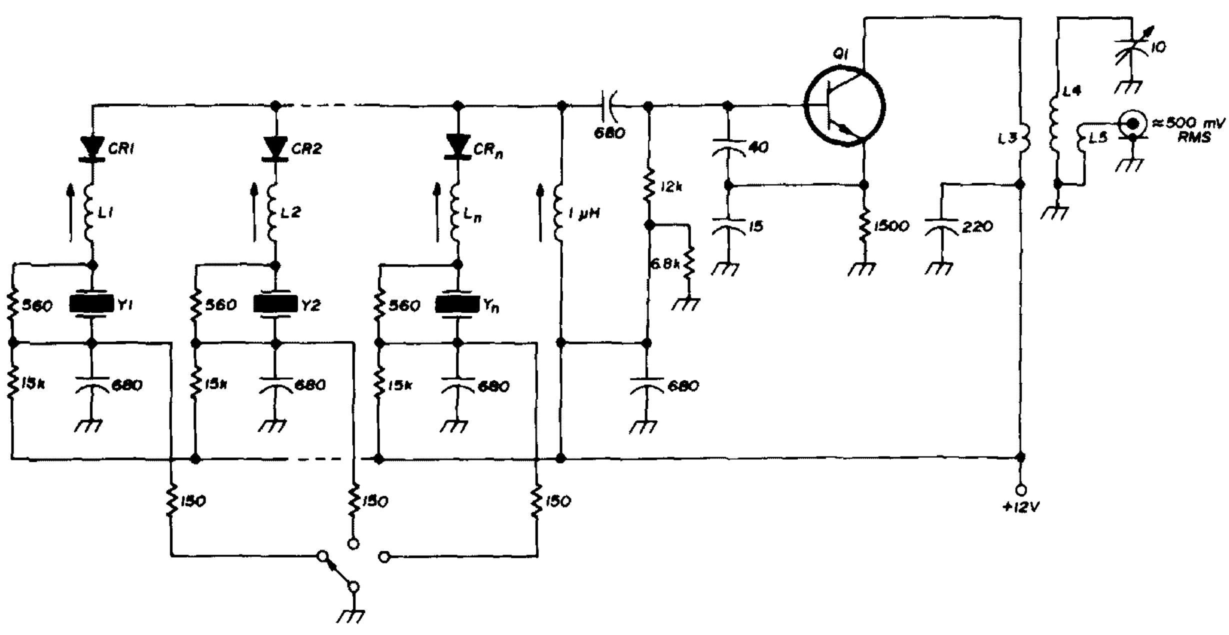


fig. 4. Switchable overtone oscillator uses third-overtone crystals in the 20 to 80 MHz range and frequency doubles in the transistor. Inductors L1, L2 thru Ln are series resonant with 10 pF at the crystal frequency. Inductor L4 in the tuned circuit at the output is resonant at the desired output frequency with 10 pF; the input and output coupling inductors, L3 and L5, have one-third the number of turns as L4. Transistor Q1 is a 2N918, BF115, HEP709 or similar. Diodes CR1, CR2 thru CRn are switching types such as the BAY67.

summary

The crystal-oscillator circuits discussed here are highly recommended for new designs since they do not require special crystal parameters. Even older crystals, which the amateur may find in his junk box, will provide extremely good results. The other major advantage of these circuits is that the influence of the circuitry surrounding the crystal is minimized.

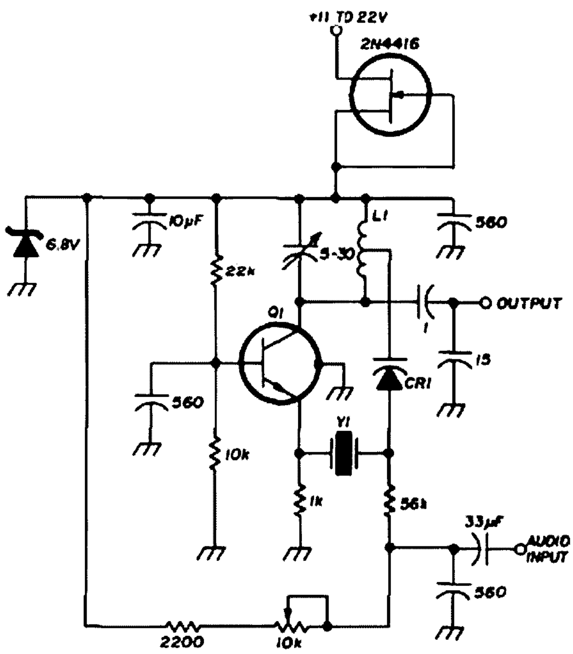


fig. 5. Overtone oscillator circuit which frequency doubles in the transistor and can be frequency modulated or used as a stable vcxo. Tuning range with a 70-MHz third-overtone crystal is typically 30 kHz at the crystal frequency (60 kHz at the output). L1 is resonant with C1 at the desired output frequency; varactor tap is at ¼ the total number of turns. Transistor Q1 is 2N918, BF115, HEP709 or similar. Varactor diode CR1 is BB142 or Motorola BB105B.

Where more expensive crystals are used, perhaps in a heated oven, the stability of these circuits will be superior to most of the oscillator circuits which are usually used in amateur equipment. This is because of the special design and the use of slightly larger capacitors across the transistor which minimize its influence on circuit stability.

ham radio

rf speech processor

for the Heath SB-102

Construction details for an rf speech processor which operates at the 3395-kHz i-f and provides up to 20-dB clipping

Timothy A. Carr, W6IVI, 452 Highland Avenue, San Mateo, California 94401

The circuit of the popular Heath SB-102 sideband transceiver, shown in **fig. 1**, lends itself to the addition of an rf speech processor. Capacitor C22 couples the output of the balanced modulator transformer, T1, to the cathode of the 6AU6 filter isolation amplifier, V2. Since C22 is mounted on the foil side of the modulator circuit board it may be easily disconnected from V2 and used to reroute the ssb signal to the speech processor. The processed ssb signal is returned to the cathode of V2. This arrangement works out very well as it does not affect any of the receiving circuitry. The 470-ohm resistor is added to the circuit so that transformer T1 is terminated with essentially the original load.

speech processor

The circuit of the rf speech processor I use with the SB-102 is shown in **fig. 2**. Although this circuit was designed specifically for use with the SB-102, it could be easily adapted to other sideband exciters which require a processor with the sideband selection filter at or near the input prior to signal processing (FL1 in **fig. 2**). The existing sideband filter is used to filter out the clipping products.

Transistors Q1, Q2 and Q3 provide the necessary signal amplification along with proper terminations for the Heath filter and automatic gain control for the two amplifier stages. The agc system is

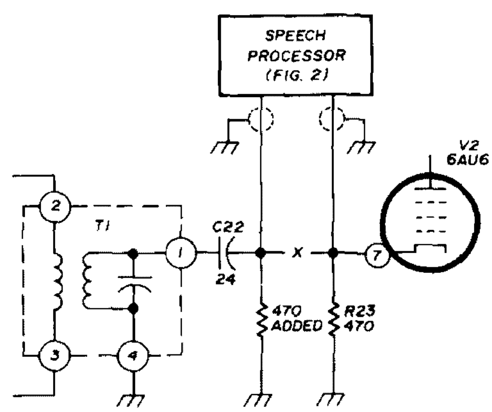
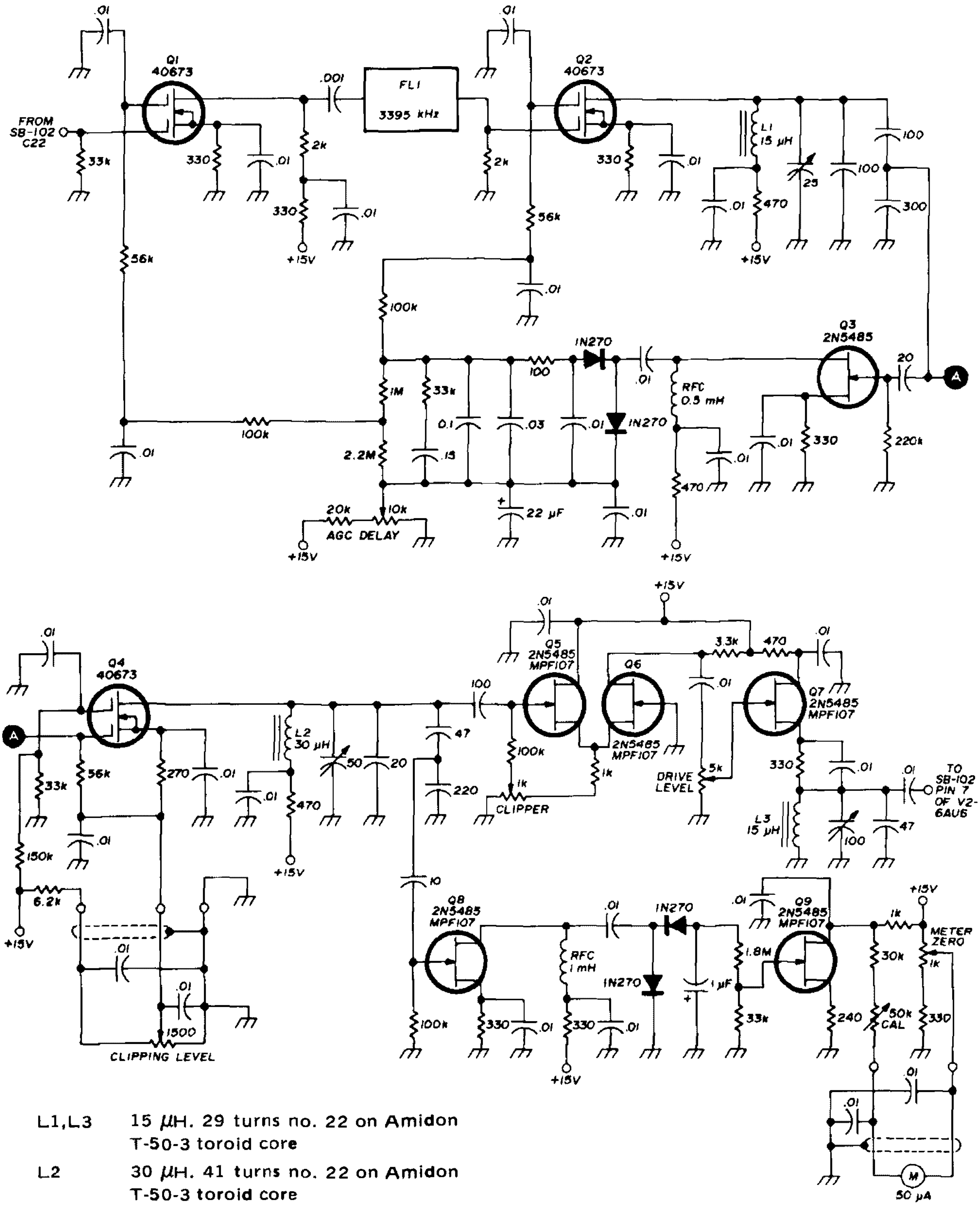


fig. 1. Rf speech processing unit for the SB-102 is inserted between the balanced modulator transformer, T1, and the filter isolation amplifier, V2.



L1,L3	15 μH. 29 turns no. 22 on Amidon T-50-3 toroid core
L2	30 μH. 41 turns no. 22 on Amidon T-50-3 toroid core

fig. 2. Circuit diagram for the rf speech processor. Filter FL1 is the same as that used in the SB-102. The existing sideband filter in the transceiver is used to filter out the clipping products.

very effective and provides a reasonably stable signal level before the actual signal processing takes place.

Transistor Q4 is a manually-controlled gain stage which allows setting of the desired clipping level, transistors Q5 and Q6 form the clipper, and Q7 is a source-follower output stage

to drive the cathode of the 6AU6. The metering circuit (Q8 and Q9) measures the relative amplitude of the signal prior to clipping; the meter is calibrated in voltage ratios (dB) above the initial clipping voltage level.

The clipper is actually a fet version of the old vacuum-tube dual-triode clipper which was popular years ago. This clipper circuit offers a high-impedance load to the preceding amplifier stage so it does not appreciably attenuate the signal prior to clipping. Gate rectification does not occur until the signal input is well above that required for 20 dB clipping. The variable tap on Q5's source resistor provides easy adjustment of clipping symmetry while Q6's drain load resistor determines the amplitude of the clipped signal.

construction

The three stages of amplification used in the processor described here result in a rather sensitive high-gain chain that is susceptible to internal feedback, oscillation and the transmitter signal pickup. When building this circuit, therefore, it's a good idea not to miniaturize the layout to the extent that you develop unwanted signal coupling between components. In addition, use toroid-cored coils to minimize inductive coupling, be liberal with bypass capacitors and decoupling resistors in the 15-volt supply line, and use plenty of shielding.

In the speech processor I built the two-stage amplifier and feedback circuit (Q1, Q2 and Q3) as well as the sideband selection filter, FL1, are located in a separate enclosure mounted toward the front panel. The gain-controlled amplifier, clipper, output source-follower and metering circuit are in another enclosure toward the rear of the chassis. All input and output connections are made through shielded cable.

The speech processor is easily connected to the transceiver. A miniature two-lug terminal strip may be installed on the mounting screw adjacent to solder point 15 (see pictorial 3-4 in the Heath SB-102 manual). Capacitor C22 may then be unsoldered from point 16 and this lead reconnected to the new terminal strip with the added 470-ohm resistor. Miniature coaxial cable (RG-174/U) may be routed from point 16 (and from the new terminal strip) through the small cutouts in the center shield adjacent to the new terminal strip — the other end of these shielded cables are connected to spare phono jacks A and B on the rear apron of the transceiver. Since no holes have to be drilled in the chassis of the transceiver, and no components are removed, the equipment can be immediately restored to original, if desired.

Kits such as the Heath SB-102 offer an economical approach to amateur radio, and this same approach was used in the construction of the speech processor circuit. More than 100 components are used in the processor, but with the exception of the 3395-kHz crystal filter, the transistors and the toroidal cores, *all components were* removed from surplus units of various kinds. Some of the component values are not optimum, but the circuit is not overly critical and there is more than enough gain available that some leeway is permissable — you can probably use the components you already have in your junkbox.

summary

Both the pre-clipping agc and the metering circuit have proven to be welcome additions to the speech processor. The unsolicited reports I have received on the quality of my ssb signal have, without exception, ranged from favorable to flattering, and have easily justified the construction of the unit.

ham radio

estimating the noise figure of your vhf system

A simple method of getting a handle on your vhf system noise figure

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The vhf and uhf mixers and rf preamplifiers used by amateurs come in assorted sizes, colors and noise figures. Some are homebrew, others are commercially made. In either case, unless you happen to be lucky enough to own or have access to laboratory type noise generators, you probably don't know positively what your vhf system noise figure really is.

The title of this article is not intended to suggest a quick way to accurately determine your particular receiving system noise figure. However, if you will take the time to read on, and then perform a few simple experiments, you can determine in what ballpark your vhf converter and preamp are playing.

You need two things to play the game: first, a signal source, preferably remotely located, providing a signal which can be picked up on the antenna and fed into the shack; and, secondly, a receiver with a well calibrated S-meter. The latter can be accomplished by inserting a step attenuator* in the i-f circuit between the converter output

*Such as that manufactured by Hewlett-Packard, Kay and others.

and the receiver input. If left in the circuit, the attenuator can be used as a calibrated i-f gain control. In any case, 1-dB accuracy is desirable.

the experiment

Fig. 1 shows a family of curves for a preamplifier with a 3-dB noise figure. The curves correspond to preamplifier gains of 4, 6 and 9 dB. *Part of the* experiment is to measure rf gain so you will know which curve to use. The curves show the signal-to-noise ratio improvement ΔS/N, with the preamplifier, in terms of mixer noise figure.

If that last statement is confusing, don't give up. First, ask yourself, "What do I *think* the noise figure of my mixer is? 8 dB?" Okay, then perform the following experiment:

1. With the antenna connected to the mixer and the receiver tuned away from the signal, set the i-f gain so the noise level registers zero S-units on the S-meter. This is the reference.* Now tune in the signal and note the S-meter reading.

2. Now add the rf stage. Note the new S-meter reading. the gain of the rf stage in dB will be the difference in the S-meter readings in dB with and without the preamp. Assume it is 6 dB.

3. Next, tune away from the signal and reset the i-f gain so the S-meter is back at the original reference setting.

4. Finally, tune the signal in again and note the final S-meter reading. The improvement in signal-to-noise ratio, ΔS/N, is the difference between this reading and that of the mixer alone. Assume it is 3 dB.

5. Locate the ΔS/N = 3 dB point on the 6-dB gain curve **(fig. 1)**. Reading down from this point, you find that you guessed quite well – the mixer noise figure is about 8.5 dB.

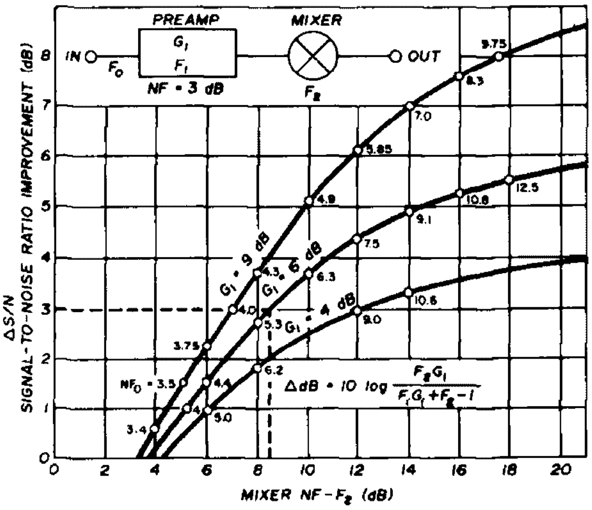

fig. 1. Overall noise-figure improvement with single preamplifier ahead of mixer (curves assume preamplifier noise figure = 3 dB). Derivation of curves is explained in appendix on page 44.

This is fine and dandy, you say, but how do I *know* the preamp noise figure is 3 dB? Very likely you don't, but you probably know what it *should* be. Assume for example that the noise figure is 6 dB, not 3 dB. Using the 6-dB noise-figure curves shown in **fig. 2**, locate the ΔS/N = 3 dB on the 6-dB gain curve. Reading down from this point, you see that the corresponding mixer noise figure is about 12 dB. If preamp noise figure is *really* 6 dB, you ought to do *something* about the mixer noise figure.

At this point you're learning that there are all sorts of interesting possibilities to this game. For example, if the mixer noise figure is truly low, say 5 dB,

*Any convenient S-meter reading can be used as the reference, but it should be close to zero so that signal readings will fall in the region below S-9 where reasonable accuracy can be obtained. Also, to perform the experiment, the receiver must have sufficient gain so that noise will produce an S-meter reading.

then signal-to-noise improvement with the preamp is harder to come by. The ΔS/N is less than 1 dB if the preamp gain is 6 dB. On the other hand, if preamp gain is 9 dB, then the noise figure improvement can be as high as 1.5 dB. But it can't be 2 dB. This is a good example of how the game is played.

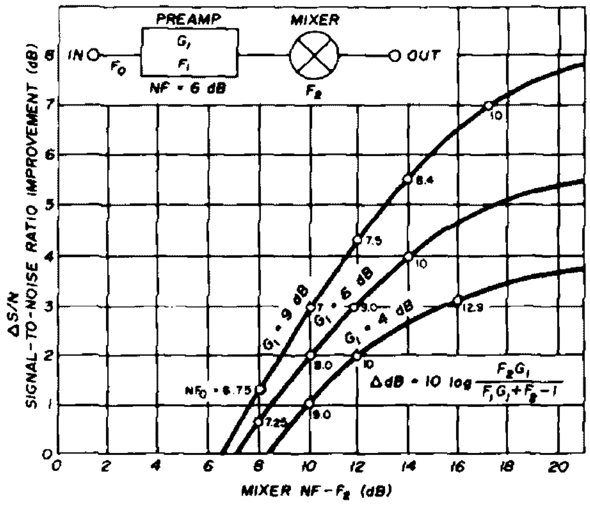

fig. 2. Overall noise-figure improvement with single preamplifier ahead of mixer (curves assume preamplifier noise figure = 6 dB).

Suppose that the improvement in S/N is negative (viz: the S-meter reading with the preamp, after resetting the noise reference, is less than the reading with the mixer alone). Your preamp noise figure is obviously greater than the mixer noise figure, regardless of what gain it has. You better do something about that, or don't use the preamp.

system noise figure

Before we end our guessing game, note that the small numbers along the curves represent overall noise figure of the mixer and rf amplifier together. These numbers are the most important of all because they represent your overall system noise figure.

Suppose ΔS/N is very large, say 8 dB, and the rf gain measures 9 dB. This means the mixer noise figure is 17.5 dB. Note that in spite of the low preamp noise figure (3 dB), the poor performance of the mixer degrades the overall noise figure to 9.75 dB!

Note also that if you use two rf amplifiers, you can still use the curves. Treat the rf amplifier feeding the mixer as the "mixer" and repeat the steps outlined above, adding the other rf amplifier to the combination.

summary

You may not be able to determine your overall receiving system noise figure accurately based solely on the instructions presented here, but if there is something seriously wrong in your system, you should be able to recognize it at once.

And remember, when you play this game, that a small ΔS/N is a sign of one of two things: a mixer noise figure almost as low as the rf stage noise figure or an rf stage noise figure almost as high as the mixer's noise figure.

appendix

The curves of fig. 1 and 2 were developed from the following equation:

$$\Delta dB = 10 \log \frac{F_2 G_1}{F_1 G_1 + F_2 - 1}$$

This equation is derived by subtracting the overall noise factor F_o from the mixer noise factor F_2 as follows:

$$F_o = F_1 + \frac{F_2 - 1}{G_1}$$

$$NF_o = 10 \log \left[F_1 + \frac{F_2 - 1}{G_1} \right] dB$$

$$NF_2 = 10 \log F_2 \ dB$$

$$\Delta NF = 10 \log F_2 - 10 \log \left[F_1 + \frac{F_2 - 1}{G_1} \right]$$

$$\Delta NF = 10 \log \left[\frac{F_2}{F_1 + \frac{F_2 - 1}{G_1}} \right] dB$$

$$\Delta NF = 10 \log \left[\frac{F_2 G_1}{F_1 G_1 + F_2 - 1} \right] dB$$

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reducing warm-up drift in the Collins S-line

Pre-heating the S-line PTO considerably reduces warm-up drift — the same technique may be applied to other equipment

Marv Gonsior, W6VFR, 418 El Adobe Place, Fullerton, California 92635

The warm-up drift of the Collins S-line may be substantially reduced by means of the relatively simple circuit addition of a two-dollar resistor, without any drilling or other circuit modifications. Applications to other receivers may be readily accomplished by experiment. All that is required is one Dale type RH-25, 25-watt, 1000-ohm resistor, or equivalent. This resistor is only about one inch (2.5mm) long, and is attached to the center of the rear vertical panel of the 70K-2 oscillator enclosure with a very small amount of cement. These resistors are designed for flat-surface, chassis mounting to assure proper heat dissipation, so they are ideally suited for the job.

The modification functions only in the receiver's *off*, inoperative mode. The resistor, R1, is wired across the power switch terminals on the receiver's control switch, S5, as shown in **fig. 1** so that a few watts of energy are heatsunk to the PTO enclosure. Once the receiver is turned on, the switch, S5, automatically shorts R1 and the PTO resumes its normal operating temperature. In the pre-heat mode, the receiver's S-meter will move slightly up from zero, and an almost imperceptible glow will be noted in the dial lamp. The power transformer will be absolutely cold. The PTO enclosure should then remain at about the same *temperature, whether or not the* receiver is in operation. The mechanical configuration for the installation of R1 is shown in **fig. 2**.

There are two alternative procedures for mounting the resistor. First, clean

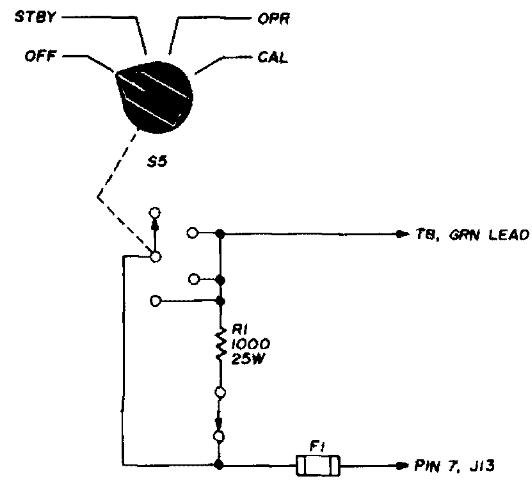

fig. 1. The 1000-ohm pre-heating resistor, R1, is connected into the circuit of the Collins receiver only when it is turned off.

the mating surfaces thoroughly with pure acetone or equivalent. Then spread a very thin film of transistor thermal compound, such as Wakefield's 120, on the resistor, taking great care to fully avoid the two mounting ears. Place a very small amount of epoxy cement in the mounting holes and on the bottom of the ears of the resistor and set it in place with a wedge against transformer

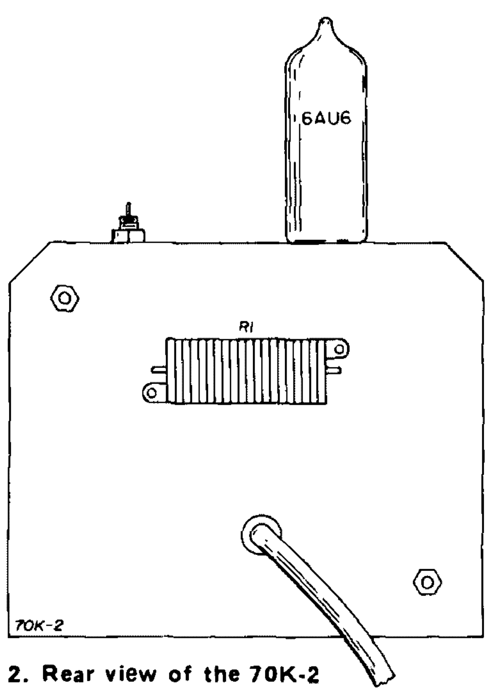

fig. 2. Rear view of the 70K-2 PTO in the Collins receiver, showing the mounting of the Dale 25-watt, 1000-ohm resistor used to stabilize the temperature of the PTO.

T3 until the epoxy cures. Hobby-type, five-minute epoxy works well.

The second method is to coat the bottom of the resistor with a very thin layer of a thermally conductive adhesive, such as Wakefield's 151-1-A, and attach it to the PTO in the same manner.

Wire the resistor according to good practice, using shrink tubing or equivalent on its terminals, since it is directly involved with the 117-volt ac line. A miniature spst switch may be mounted on the power switch and wired in series with the resistor to inactivate it during

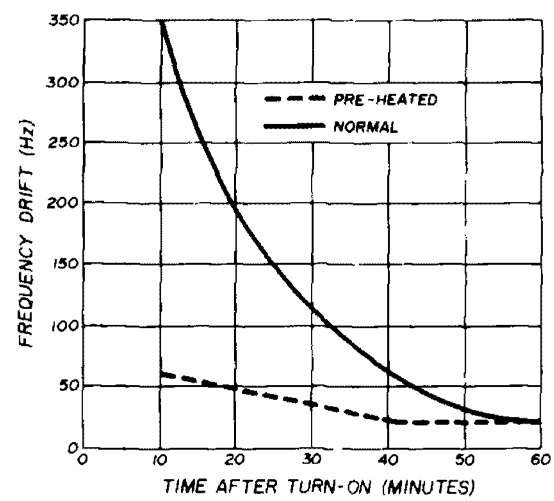

fig. 3. Warm-up drift of the Collins S-line is reduced considerably when the PTO is pre-heated by the added power resistor. This data taken at an ambient temperature of 20°C.

extended standby periods or, alternatively, the power cord may be unplugged.

The frequency measurement data at 14 MHz were taken on a H-P model 5381A, a seven-digit frequency counter in combination with a precision external time base. This was used in conjunction with a Rec-Counter.[1] The Δ_f curves shown in fig. 3 reveal the relative improvement in my 75S3-B; a worthwhile four-to-one warm-up drift reduction, in the first hour, for chasing DX on cold winter mornings.

reference

1. K. Macleish, "The Rec-Counter," *QST*, May, 1971, page 11.

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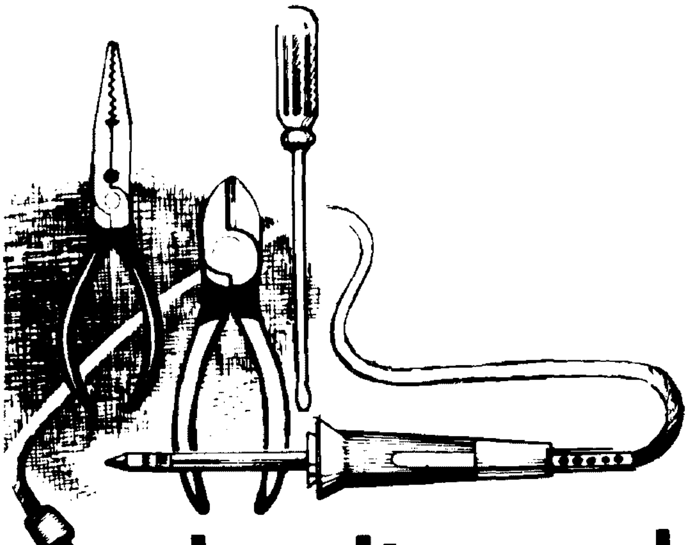

circuits and techniques

ed noll, W3FQJ

experiments and projects — cosmos

In the months ahead this column will be devoted primarily to two subjects: power generation using solar energy and wind, and new solid-state technology. My plans for generating electrical power with solar energy and wind have been previously discussed in this column. A 200-watt wind generator has been ordered, and by the time you are reading this I hope to have the system in operation. The amateur station and workshop at W3FQJ will be powered completely by solar and wind energies.

The second major subject, solid state, will include discussions of bipolar, fet, mosfet, integrated-circuit and cos/mos technologies. Electronics, more than many other industries, has always been interested in the conservation of electrical energy. Solid-state devices are a revealing example. Their low operating voltages and lack of current-demanding filaments and high-voltage power supplies make them ideal for powering with the non-polluting sun and wind. Radio amateurs are to be commended for their leadership in these endeavors. In fact, their enthusiasm for QRP operation and the use of minimum power in the sustaining of any QSO are exemplary.

Many amateurs would like to take these devices on mentally, but have procrastinated or take little pleasure from learning the theory of a device without doing some practical experimentation. While it is true that many amateurs have built complex solid-state devices following detailed construction information supplied with kits or in magazine articles, what happens within the device and circuit is often vague. There are a good number of amateurs who, as yet, have not built their first solid-state stage. Perhaps this column will be able to lead you to a better understanding of device function and circuit.

The plan is to combine individual projects with experimental steps (which I call *expros*). First you will experiment with the device, learning about its innards and external operating characteristics. Then you will build a project

using the one or several of the devices. **Expro 1** will be the jfet and will conclude with the construction of a two-stage QRP transmitter. Most expros will be built on perf boards which, if desired, can be used over and over again.

cosmos logic

In previous columns there have been basic presentations on the fabrication and general operation of most solid-state devices. One which was overlooked was the cosmos. Fundamentally, the cosmos device incorporates enhancement-mode mosfets into integrated circuits. Uniformity, balance, stability, versatility, reliability and compact size are the reward of IC fabrication. The special star of the cosmos device is its conservation of power.

The enhancement mode of mosfet operation, integrated fabrication and complementary symmetry circuitry combine to form the increasingly popular RCA series of cosmos integrated circuits. These devices require very little power (standby power measured in microwatts), operate over a wide supply

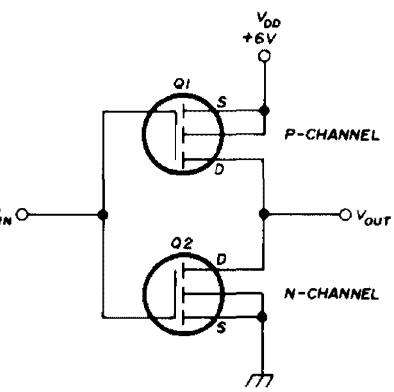

fig. 1. Two enhancement-mode mosfets in complementary symmetry.

voltage range (less than 2 to more than 15 volts), a wide temperature range and have the high input impedance characteristic of all mosfet devices. Cosmos devices are ideally suited to logical, digital and switching, as well as linear applications. Typical package dissipation is 200 milliwatts.

What is complementary symmetry? It is the basic circuit that makes cosmos tick and be so conservative in power demand. It is the jewel of the modern watch.

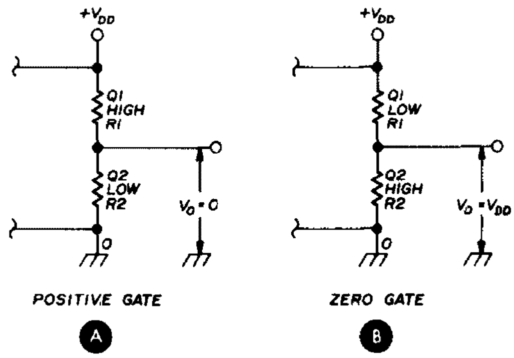

fig. 2. Equivalent operation of a complementary-symmetry pair (see text).

Field-effect transistors come in two forms, n-channel and p-channel. One type is the complement of the other. The channel charges move in opposite directions. When two such devices are connected in parallel, **fig. 1**, unusual operating conditions arise. The top transistor is a p-channel enhancement-mode mosfet; the bottom transistor, an n-channel type. A positive voltage applied to the parallel gates causes the n-channel device to conduct. The p-channel unit remains off.

What is the output of the complementary symmetry stage? Assume that the gate voltage is made positive by the amount of the supply voltage V_{DD} (6 volts). The lower transistor, Q2, conducts, presenting a low-resistance path (perhaps one to several thousand ohms) between the drain and common. The top transistor, Q1, is cutoff. The resistance of its path is in the thousands of millions of ohms. Under this condition what is the output voltage, assuming that the load placed on the output is the

gate circuit of a second high-impedance enhancement-mode mosfet?

As shown in the equivalent circuit, **fig. 2**, a two-resistor voltage divider is set up. Inasmuch as the top resistance is

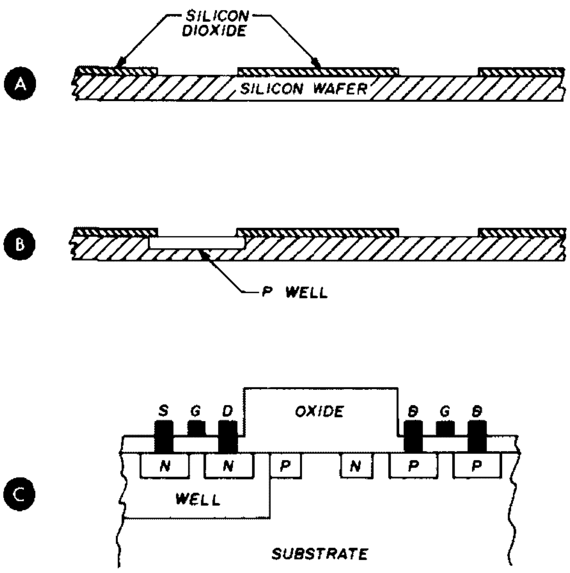

fig. 3. Basic construction of a mos integrated circuit.

very high in comparison to the lower one, there is practically no voltage drop across the lower one and the output is in effect zero volts (ground or common potential). Furthermore, very little current is drawn from the supply line because of the extremely high resistance of the series combination of R1 and R2.

If the gate voltage is zero (common potential), transistor Q1 conducts while a cutoff bias is applied to transistor Q2. Why does transistor Q1 conduct with zero voltage applied to the gate? Note that its source is at positive potential so the gate of the p-channel device is negative in comparison to the source; therefore, it conducts. In **fig. 2B** the upper transistor presents an approximate 1000-ohm path between V_{DD} and output, while the path between the output and common is in the millions of ohms.

Now the high value resistance is in the path between output and common; the low value between the supply voltage and output. Consequently the output voltage is positive and essentially the same value as V_{DD}. Again the summation of the two resistances is extremely high and there is little current demand made from the supply source.

In summary, swinging the gate voltage between the supply voltage value and zero causes the output to change between zero and the supply voltage value. This is an ideal situation for digital and switching applications.

It should be noted that in both cases the resistance of the path between supply voltage V_{DD} and common (assuming no low-resistance load is placed on the output) is in the millions of ohms. Therefore, for both steady-state conditions, the resting power demand is exceedingly low. This is great for logic circuits. Logical zero and logical 1 states make the same low power demand on the supply.

In fact, the only time that any significant power is drawn occurs when the gate voltage is in transition between the two steady states. Furthermore, the higher the speed of the transition (rate of rise and fall of the leading edges), the lower the power demand made on the

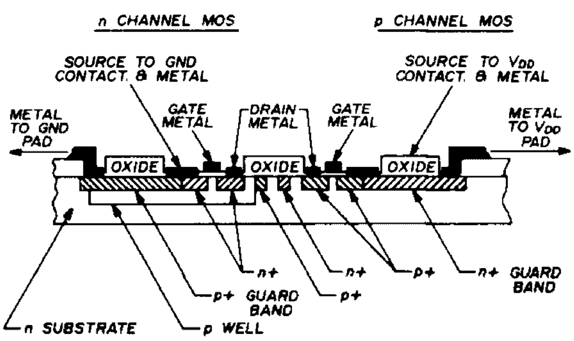

fig. 4. Fabrication plan of RCA cosmos devices.

supply. This is to state that the power demand will be less when an incoming signal is a steep-sided pulse or square wave; more if the incoming signal is a sine wave.

The complementary symmetry circuit is basic to the cosmos device. Why the need for integration? In practice it is difficult to obtain exact symmetry between n- and p-channels. The greatest advantage of the complementary symmetry connection can be made when it is applied to integrated-circuit technology. In the single monolithic chip, uniform and balanced channels can be processed using the diffusion procedure.

A basic RCA cosmos device begins with a silicon substrate and silicon dioxide islands that have been deposited on its top surface using heat deposition and photolithographic procedures, **fig. 3**. Diffusion steps are then used to form the various elements and isolating barriers. When using the n-type substrate it is necessary to first diffuse a p-type well in that substrate to serve as the base needed in the formation of the n-channel transistor. The basic makeup of the two-transistor complementary devices is shown in **fig. 3C**. On the left, in its well, is the n-channel unit. On the right, diffused directly into the substrate, is the p-channel unit of the complementary pair.

The complete RCA package arrangement is shown in **fig. 4**. In addition to the complementary stages there are guard bands which surround and protect the separate mos devices, well, diodes,

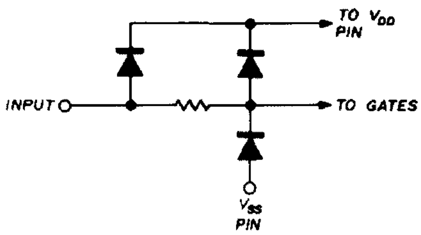

fig. 5. Protective input diodes.

etcetera. They provide isolation and prevent leakage. Guard bands also provide conduction paths to the external supply voltages.

Included as a part of the fabrication are protective diode systems. The input diode arrangement, **fig. 5**, protects the device from static charges and input voltage transients. This diode clamping keeps the device and extraneous volt-

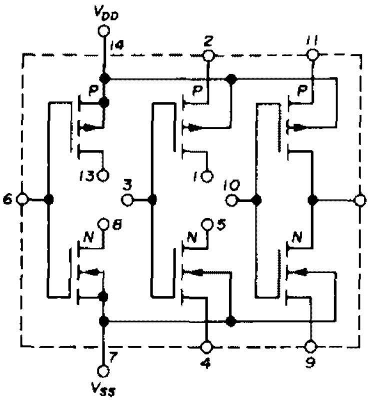

fig. 6. RCA CD4007A cosmos IC consists of two complementary-symmetry stages plus an inverter.

ages at safe levels. Nonetheless, the device must be handled carefully in accordance with the usual mosfet precautions.

The complementary symmetry circuit, when connected as shown in **fig. 1** and included in the cosmos integrated circuit, is called an inverter. As mentioned, a positive voltage applied to its gate results in a decrease in the output voltage. A positive change at the input results in a negative change at the output.

In terms of logic language, a logical 1 input (+ voltage) causes a logical zero output. A logic inversion has taken place. Conversely, with the logical zero input, there is logical 1 output.

Three complementary symmetry circuits are built into the RCA CD4007A cosmos integrated circuit, **fig. 6**. The first two configurations are referred to as complementary symmetry circuits. Note that separate drain terminals are brought out (pins 1, 5, 8 and 13). This provides versatility in interconnecting the two stages into various forms of complementary-symmetry-stages. The third stage is the basic inverter and its

circuit is identical to that of **fig. 1**. It differs only in that the two drains are connected together internally at pin 12.

One complementary pair and the inverter can be connected into a monostable multivibrator as shown in **fig. 7**. In a resting state the p-channel mosfet of the stage on the left is biased off; the n-channel on. As a result the C output is low and the inverter D output is high because the output of the first stage is connected to the gate of the second.

When a negative pulse is applied by way of capacitor C1, the p-channel mosfet is turned on and the n-channel off. Capacitor C2 begins to charge to the supply voltage and this positive voltage on the gate of the inverter drives output D low. This happens very quickly aided by the feedback path between output D and the input of the n-channel section of the input pair, eventually shutting off the n-channel of the first stage.

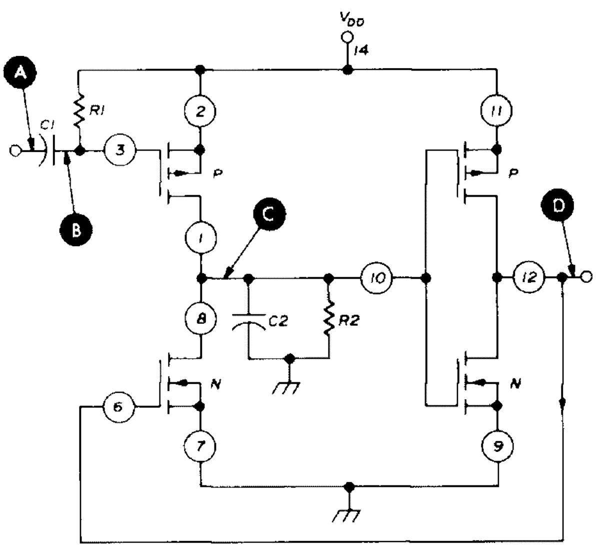

fig. 7. Using the RCA CD4007A cosmos IC as a multivibrator. Operation of this circuit is described in the text.

Capacitor C1 now begins to charge to the supply voltage, V_{DD}, through resistor R1. The p-channel mosfet remains on until the charge interval is such that the p-channel device is shut off by the declining negative voltage on capacitor C1. However, the n-channel device of the first inverter also remains off until capacitor C2 discharges sufficiently through resistor R2 toward common.

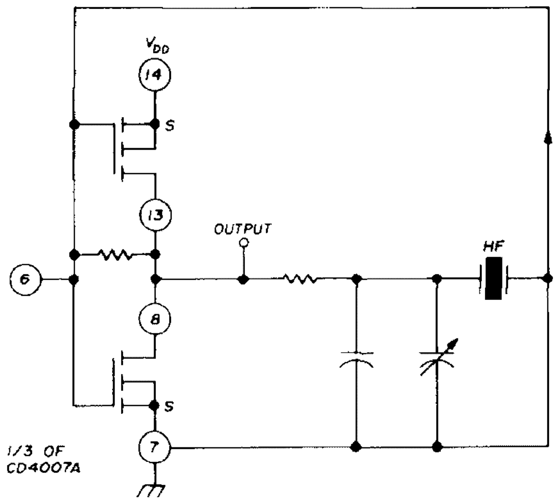

fig. 8. High-frequency crystal oscillator circuit is based on two stages of an RCA CD4007A IC.

When the turnover or threshold voltage of the p-channel transistor of the inverter is reached, the device is turned on. The output voltage at D then begins to rise and the n-channel device of the first stage is switched on, providing a low resistance discharge for capacitor C2. This causes the operation of the first stage to go from high to low and the cycle is completed. The multivibrator then remains in its resting state until another negative pulse arrives at point A.

The circuit of **fig. 8** shows how a single complementary pair can be connected as a high-frequency crystal oscillator. The output is taken at the common drain circuit. The feedback arrangement is Pierce-like in the form of a lowpass crystal filter that is connected between the drain circuit and gate input. The small trimmer capacitor permits fine adjustment of the crystal oscillator frequency.

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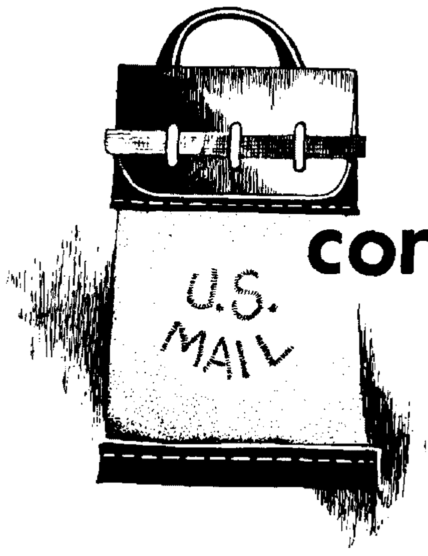


comments

Dear HR:

While interest in speech processing is understandably high, I wonder why the same degree of interest is not shown for the CW mode. Perhaps as amateurs we are convinced we know all there is to know. A little research can, however, raise some interesting questions. Consider the question of minimum bandwidth requirements which are expressed by Shannon's formula in terms of bits per second per Hz of bandwidth as

$$\log_2 (1 + s/n)$$

where s/n is the signal-to-noise ratio. For 50-bit words and Morse code at 25 wpm the bit rate is close to 20 bits per second. With an arbitrary signal-to-noise ratio of 15 the formula gives 4 bits per second per Hz, or 5-Hz bandwidth for this example.

Unfortunately, we cannot achieve this signaling rate in practice, and for a two-state amplitude-modulated signal such as CW the bandwidth requirements would be approximately three times the theoretical minimum (15 Hz in this case). Since bandpass audio filters with this degree of selectivity are easily built, why are filters of 100 and 150 Hz in common use, and filters of 50 Hz or less very unusual?

One reason is that some CW operators rely almost exclusively on their personal selectivity built into their ears and brain. I believe there is a lot to learn in this area, but in my own particular case, the better the signal-to-noise ratio reaching my ears, the better I copy. The most common reason for the rarity of minimum-bandwidth filters is the phenomenon of filter ringing. I cannot recall seeing a single article on this aspect in the amateur magazines, and the purpose of this letter is to stimulate some correspondence on this subject.

My own knowledge is inadequate, but it appears that filter ringing arises basically from two effects. First, the filter's group-delay characteristic, and second, the filter's transient response. From experience with active bandpass filters I believe that the delay characteristic can be adjusted for minimum distortion of the keyed audio envelope so bandwidths of about 50 Hz can then be used. The question of transient response remains, and at the moment I don't

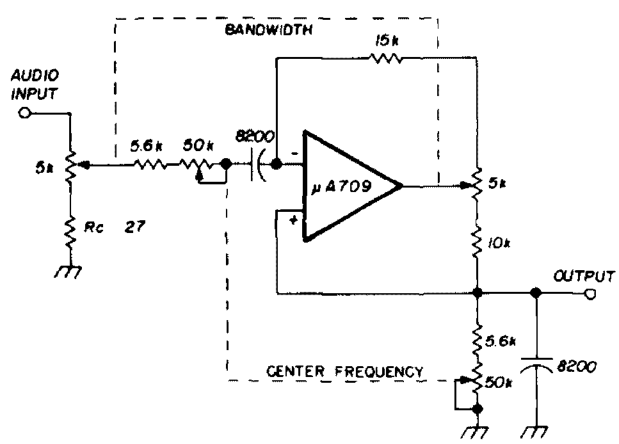


fig. 1. Variable-bandwidth, variable-frequency audio filter. With circuit values shown center frequency can be tuned from 300 Hz to 3 kHz. (A 741 op amp would be somewhat better over this frequency range.)

have an answer, although I have hopes of minimizing this, too.

In the commercial world of data transmission many signaling modes other than two-state a-m have been evaluated. Generally they succeed in trading signaling rate for immunity to interference. Some of these signaling modes may well have amateur applications. Also, the technique of coherent detection seems promising. The question of spectrum shaping has been studied and optimum relationships between transmitter and receiver filter characteristics have been established. Perhaps in our case we should pay more attention to the optimum design of key-click filters and receiver audio filters.

For anyone who has not yet tried a variable-frequency, variable-bandwidth audio filter, I would recommend the circuit shown in fig. 1. Or, more simply, the IC package made by Kinetic Technology, Inc.* Their FX60 active bandpass filter requires only one fixed and two variable external resistors and costs about $6.00.

Ron Skelton, 6Y5SR
Kingston, Jamaica

memory keyer mods

Dear HR:

I found in my memory keyer that the 1101A random-access memories (used in place of the 25L01s) would play back a message erratically. This can be cured by placing a 1000 ohm, 1/4-watt resistor from the *data output* (pin 12) of the 1101 to ground. Note that pin 12 of memory A is connected to pin 12 of memory B. The resistor can be soldered to the foil side of the PC board. Put insulating tubing on the resistor leads to prevent the leads from shorting to the foil.

A feature which will increase the versatility of the keyer when using it as a straight electronic key without a memory is to replace the *memory select* switch, S3, with a center-off spdt switch. Placing S3 in the center position will not allow either memory to send an output to the keying transistor. The keyer can then be used as a normal electronic key without worry that the memories will inadvertently trigger and send an output to the keying transistor.

I would like to thank my friend Bob Thing, WA3WKJ, for the ideas presented here.

Sam Guccione, W3GVP
Camden, Delaware

*Kinetic Technology, Inc., 3393 De La Cruz Boulevard, Santa Clara, California 95051.

information needed to update RSGB history

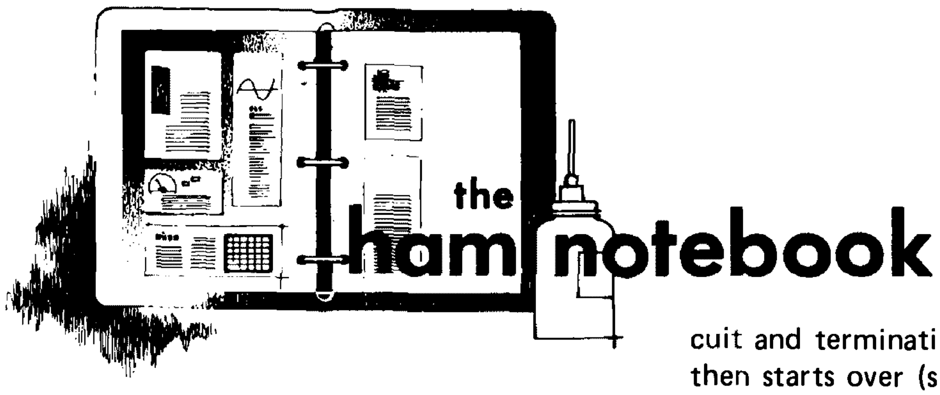
Dear HR:

Many readers of *ham radio* are also members of the Radio Society of Great Britain, and are familiar with the book, *World at Their Finger Tips*. This book was written by the late John Clarricoats and covers the history of the society and the work of many of its members from 1913 to 1963.

The RSGB has honored me with the task of writing a sequel to this book in order to bring the society's historical records up to date. During the past decade the RSGB and numerous members have contributed to the tremendous advance in all fields of radio communications throughout the world.

In order for me to make a success of this book, and do the society justice, I must have information, therefore I appeal to RSGB members who read *ham radio* to send me details of their radio achievements during the past ten years. I would like to have this information as quickly as possible because there is a lot to do, and I hope to have the work completed within a couple of years.

Ron Ham
Faraday, Greyfriars
Storrington, Sussex
RH20 4HE England



fm equipment interface problems

In trying to use two current pieces of fm equipment, a 10-watt transmitter (Regency HR-2) and a 100-watt amplifier (Dycom), it was found that the two pieces would not work together. To begin with, the transmitter is well designed and has excessive standing wave protection — a very good move, especially if the wrong antenna is inadvertently connected, or left disconnected. This safety device will probably save replacement of expensive power transistors.

The amplifier has an "automatic switching" feature, meaning that dc power is applied at all times, and as rf excitation is applied, the unit turns itself on. When the transmitter is turned on (microphone button pressed), rf appears across diode CR1 (see **fig. 1**), and a dc voltage appears at point A. The rf choke isolates the rf voltage at point A and conducts dc to the base of transistor Q4, placing a positive bias on the base, and transistor Q4 turns on, picking up relays K1 and K2. Unfortunately, the moment K1 and K2 start to transfer, the back contacts open, opening up the line from the transmitter, and the swr protection circuit in the exciter cuts in, shutting off the power. The relays then open, remaking the exciter output circuit and terminating the line. The cycle then starts over (since the transmitter is still on). The relays just sit and vibrate like an old fashioned door bell — a revolting state of affairs.

The fix is ridiculously easy. Just connect a 500-μF capacitor (12 volts or more) between the base of Q4 and ground. As diode CR1 conducts, the 500-μF capacitor, C4, is charged and holds its charge long enough for the relays to pull completely in. Once the relays are picked up, the line from the transmitter is terminated by C2-C3-L1, usually with less than 1.2:1 vswr, particularly if adjusted carefully.

Capacitor C1 series resonates with the interconnecting leads and relay elements to allow maximum receiver performance and proper transmitter operation when the amplifier is not in use. Power may be switched from the 100-watt amplifier to low power only by opening the circuit at the fuse.

The changes suggested in these notes should make the use of the equipment much more satisfactory and enjoyable.

Dave Chapman, W9DPY

Heath SB102 modifications

I have two Heath SB102 transceivers and both have shown the same two difficulties, which appear to be generic: First, objectionable audio hum level,

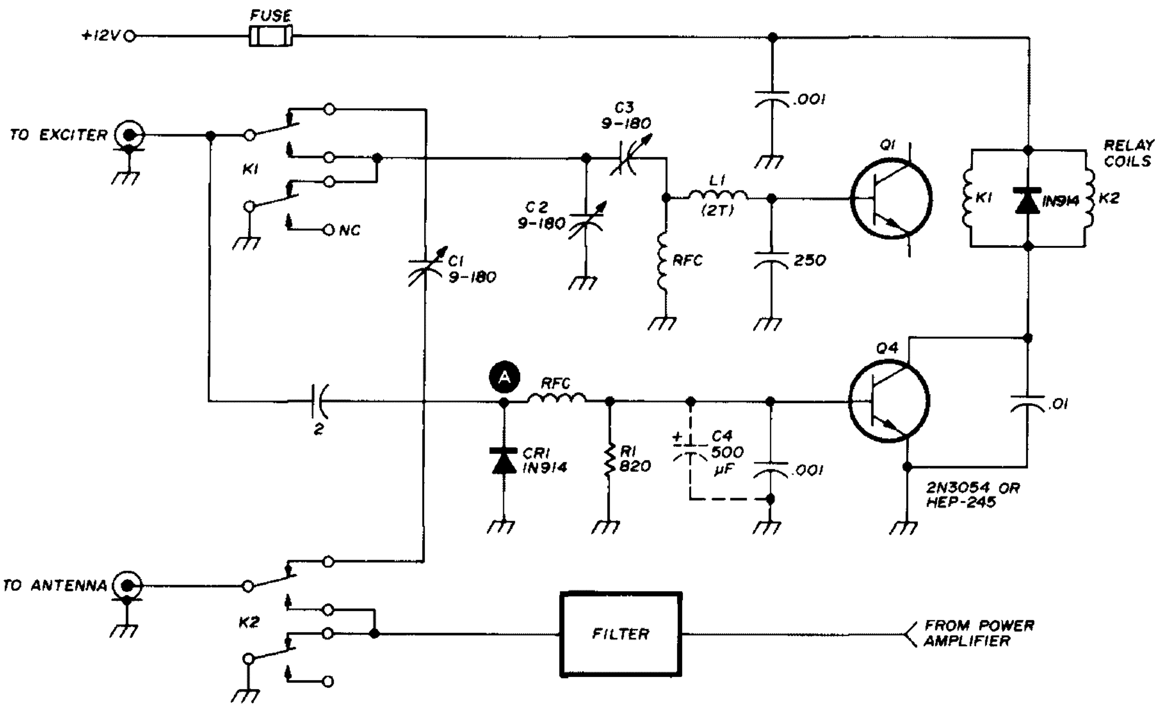


fig. 1. Added 500-µF capacitor allows automatically switching relays to be picked completely up before the swr protection circuit drops them out.

and secondly, one resistor that runs hotter than it should. As for the hum, an additional filter did not help, and shorting the arm of the volume control had no effect. As a part of the cable harness there is a shielded lead from the arm of the volume control to capacitor C308 which couples the signal into the grid of V14A, the audio amplifier. Replacing this lead with a *separate* shielded lead eliminates the objectionable hum level.

Resistor R955 is a 100k, ½-watt resistor which avalanches down in value and burns up. Replacing it with a 1 or 2 watt, 100k resistor clears up this problem.

Lowell White, W2CNQ

zener-diode noise

Recently I was asked to convert some vhf preamplifiers and replace the original 417A tube with a low-noise transistor. This required the addition of a zener diode to provide the correct voltage for the transistor. The zener was originally mounted very close to the low-noise transistor and a noise generator was used to determine the noise figure. The results were nearly the same as when the original tube was in use. The zener was then mounted away from the transistor, near the power source. A noise figure measurement indicated a noise reduction of 2 dB.

Vern Epp, VE7ABK

short circuit

The 67-pF ceramic capacitors used in the lowpass filter described in the March, 1975, issue of *ham radio* have all been sold. An alternate capacitor is the 140-pF APC capacitor available for 50 cents from CPO Surplus, Box 189, Braintree, Massachusetts 02184. Remove one plate for 136 pF, fully-meshed capacitance. Since CPO Surplus has a handling charge of 50 cents on orders less than $3.00, the two capacitors for the lowpass filter can be obtained for $1.50, postpaid.

a second look (from page 4)

table 1. Partial listing of power amplifier tubes in current use on the amateur bands.

tube type	number	total filament/heater power (watts)	equipment manufacturer
thoriated tungsten filament			
572B	2	51	Heathkit
811	4	100	Collins 30L1
813	2	100	
833A	2	200	
3-400Z	2	140	Henry
3-500Z	2	140	Drake, Heath
3-1000Z	1	157	BTI
3CX1000A7	1	152	
3CV1500A7	1	152	
4-125A	2	65	
4-250A	2	140	
4-400A	2	140	E. F. Johnson
4-500A	2	204	
4-1000A	1	157	
4CX1500A	1	200	Henry
indirectly-heated oxide cathode			
8072	1	17	CX-7A
8122	2	35	National
8873	2	40	Heath, Henry
8874	2 or 3	40 or 60	ETO, Henry
8877	1	50	ETO, Henry
4CX1000A	1	57	Collins 30S1
4CX1500B	1	60	

desirable. If the 6-dB differences proposed in Docket 20282 are used, then one-quarter and one-eighth, respectively, of the above emitter wattages would coincide.

If a filament power limitation for each class of license were to be adopted, Quinn has suggested that no input or output power limitations be imposed upon the amateur service. The filament power limitation would predetermine maximum operation conditions as the tube could only be driven up into plate current saturation. The individual operator could exercise his own initiative and technical ability, using any tube which fell within the authorized emitter power limits. This is the same sort of initiative which prompts some amateurs to build large antenna systems to enhance their signal strength, and adds to the competitive spirit of the hobby and advances

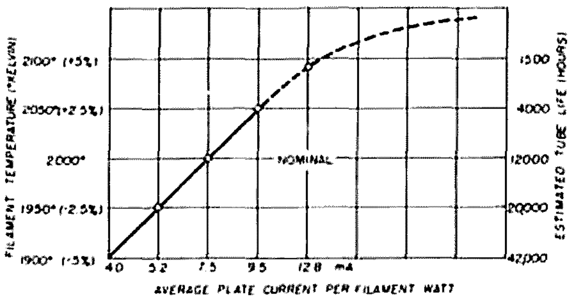


fig. 1. Approximate comparison of filament temperature and emission life vs average plate current (mA) per filament watt for thoriated-tungsten power amplifier tubes. Note that if one or more maximum tube ratings are used simultaneously tube life is severely decreased.

the technical achievements in amateur radio.

The establishment of total emitter power for amateur linear amplifiers would automatically establish a maximum power, as shown in figs. 1 and 2. The dotted regions shown on these curves represent diode operating conditions only. In actual practice, if a tube were operated in this area under rf conditions, it would probably fail within a few hours due to either control or screen grid failure, excessive internal anode temperature, or oxide cathode evaporation, etc. In view of this, the estimated tube life shown would not be representative at the upper current levels because the tube would probably fail catastrophically, rather than from loss of emission.

There will, of course, be those opera-

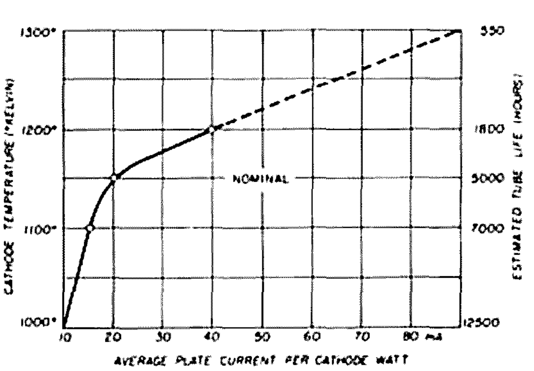


fig. 2. Approximate comparison of cathode temperature and emission life vs plate current (mA) per cathode watt for power amplifier tubes with indirectly-heated cathodes. Note that if one or more maximum tube ratings are used simultaneously tube life is severely decreased.

tors who will obtain a few more watts and minimize short tube life by reducing filament power during standby, and increase filament power during rf drive conditions, or increase plate voltage, to maximize plate efficiencies. However, as the old saying goes, "You can't get something for nothing," and equipment and tube manufacturers can tell upon inspection if their product has been abused. One or two dB would not be worth the effort.

Presently, it is very difficult, if not impossible, for the FCC to monitor and police amateur power limits. If the technique suggested by W6MZ were adopted, and an amateur was found to be using a power amplifier which used tubes with manufacturer's filament/heater ratings in excess of the maximum specified wattages, it would be a simple black or white infringement of the rules.

Jim Fisk, W1DTY
editor-in-chief

new products

electronic keyer

Data Signal recently introduced a new printed-circuit electronic keyer which is offered in two versions: one for TTL, the other for CMOS logic. The keyer is complete, including potentiometers, a large speaker and all mounting hardware. The user must supply an enclosure, a keying paddle and a small 5-volt power supply for the TTL version or a 9-volt transistor battery for the CMOS version.

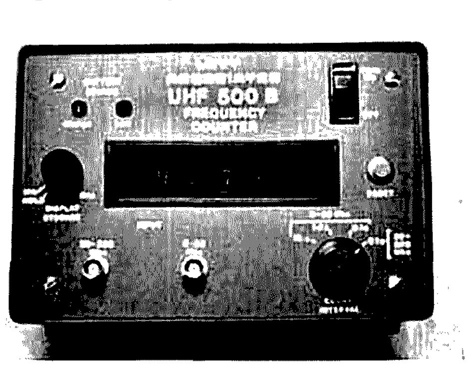
The keyer circuit includes fully automatic, self-completing dots, dashes and spaces. Each dot and dash is provided with its own jam-proof space, eliminating any chance of jamming dots and dashes together. At the instant the paddle is closed, the start-stop oscillator starts sending the code element. Key closures extending to the end of the dot-dash space are allowed with assurance of no additional dot or dash. This wide keying tolerance makes this keyer extremely easy to use, for beginners and old timers alike. A variable weight ratio control is provided to allow the operator to adjust the dot-dash to space ratio of each character. The built-in audio system includes a full-range audio oscillator, volume control, tone control and speaker. Keying speed is adjustable from 5 to 50 wpm.

These keyers eliminate the number-one source of keyer failure — the output keying relay. The reed relays used in many electronic keyers are subject to pitting, sticking and breakage. In the Data Signal unit output keying is accomplished with specially selected high voltage, high current transistors. These heavy-duty transistors require a small amount of current for operation and are designed to handle the two most often used keying systems — grid block keying and solid-state transmitters.

The two versions of the keyer, TTL and CMOS, allow the user to select the keyer best suited to his own requirements. The TTL version is ideally suited for home stations where a common 5-volt supply is available, while the CMOS version requires very low current and is just right for QRP or portable operation.

The TTL version of the electronic keyer, model TTL/PCK-1, is priced at \$19.95 wired (\$14.95 kit); the CMOS version, model CMOS/PCK-1, is \$24.95 wired (\$19.95 kit). For more information, write to Data Signal, Inc., 2212 Palmyra Road, Albany, Georgia 31701, or use *check-off* on page 94.

500 MHz frequency counter



The new UHF 500B frequency counter from Levy Associates features laboratory accuracy in a portable instrument

More Details? CHECK—OFF Page 94

at an inexpensive price. This counter, which uses the latest state-of-the-art advances, provides all of the features of instruments costing three times as much: built-in nicad battery with charger for easy portability, 7-digit display, high sensitivity, high-stability temperature-compensated crystal oscillator, and response to at least 500 MHz (575 MHz typical). The bright LED displays and polarizing filter make reading easy in high ambient light conditions. Adjustable display storage provides for minimum transmitter *on* time.

The sensitivity of the new UHF 500B frequency counter is 35 mV or less to 50 MHz; less than 150 mV from 50 to 500 MHz. With the uhf preamp option ($85) the sensitivity is 30 mV or less to 500 MHz. Other options include 1000 MHz capability ($195), high-stability time base with ±0.0001% accuracy from zero to 40°C ($50), and green or yellow displays [red is standard] ($15). The standard UHF 500B is priced at $525. For more information, use *check-off* on page 94 or write to Levy Associates, Post Office Box 961, Temple City, California 91780.

specialized communications techniques for the radio amateur

This new book from the ARRL was written by those experienced in each of the fields which it covers. The seven chapters provide practical details on communications techniques, amateur television, slow-scan television, facsimile, RTTY, space communications and advanced techniques.

The chapter on amateur television contains circuit details and applications information for television cameras, transmitters and receiving techniques. The section devoted to facsimile in-

cludes conversion of Telefax transceivers and reception of weather-satellite pictures. Radioteletype is thoroughly covered with sections on equipment and hardware, AFSK, RTTY measurements, and much more. Moonbounce (EME), satellites, laser fundamentals and communications and digital communications are covered in the chapter on space communications and advanced techniques. Softbound, 208 pages, $3.00 from Ham Radio Books, Greenville, New Hampshire 03048.

gutter clamp mobile antenna

Answering the need for quick-on, quick-off mobile antenna mount is the Larsen Gutter Clamp Antenna Mount that goes on or off the vehicle faster than you can tell about it. Styles of the clamp are available to fit all popular makes of antennas. Complete mounts and antennas of both the quarter-wave type and gain antennas for low, high and vhf bands, complete with the clamp, are available.

Finished in weather-resistant irridium, the Larsen Gutter Clamp features a high-tension spring that clamps the unit to any make of vehicle. A

More Details? CHECK—OFF Page 94

special rubber bumper prevents marring the car's finish. Provision is also made to assure a proper ground plane for maximum radiation efficiency of the antenna. The Larsen Gutter Clamp Mount comes complete with ten feet (3m) of coax and plug already installed.

For more details regarding the Larsen Gutter Clamp Mobile Antenna Mount write to Larsen Antennas, Post Office Box 1696, Vancouver, Washington 98663, or use *check-off* on page 94.

linear ic handbook

A new *Linear Integrated Circuits* handbook is now available free of charge from National Semiconductor Corp. The handbook gives complete descriptions and specifications of 280 linear IC devices that are useful in building nearly all types of electronic systems, ranging from communications and consumer-oriented circuits to precision instrumentation and computer designs. It covers product-line categories that include voltage regulators, operational amplifiers, voltage comparators and buffers, functional blocks (timer circuits), ICs for consumer products, transistor and diode arrays, and analog switches.

Most of the product descriptions are accompanied by application advice, and the text is supported by graphs, charts and diagrams. Selection guides are provided for operational amplifiers, fixed-voltage regulators, variable-voltage regulators, and voltage comparators. Cross reference guides are provided for fet-input op amps and the linear products of Texas Instruments, Motorola, Signetics and Fairchild.

For a free copy of the *Linear Integrated Circuits* handbook, send a postcard to the Marketing Services Department, National Semiconductor Corp., 2900 Semiconductor Drive, Santa Clara California 95051, or use *check-off* on page 94.

$1.00

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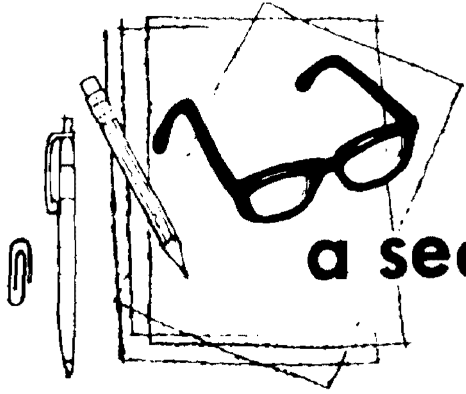


a second look

by jim fisk

In our modern day world of solid-state electronic gadgets and centralized urban living, it's the rare amateur who hasn't been troubled at one time or another by interference complaints. As often as not the interference is being generated by some other source, but if you have a tower in your backyard, you're a likely suspect and the first one to whom they turn when the local taxicabs (or whatever) come booming through their quadraphonic stereo systems.

As I pointed out in this column last year, the problem can be effectively cured only by proper design and construction at the manufacturing level. The home-entertainment business is highly competitive, however, and with today's staggering economy the manufacturers are reluctant to add filtering and lead bypassing that would increase the price tags on their equipment. Until recently, in fact, they contended that only 1% of home entertainment equipment operates in an rf environment which necessitates special attention. However, with the proliferation of two-way radio systems, as well as higher power a-m and fm broadcasting stations and high-speed digital systems which can cause interference, I doubt that many consumers would agree.

The answer to this problem may now be in sight. On May 15th, Rep. Charles A. Vanik of Ohio introduced a Bill to the House of Representatives which, if it becomes law, will give the FCC the right to regulate the manufacture of electronic home-entertainment devices to reduce their susceptibility to interference from nearby radio transmitters.

That Bill, H.R. 7052, has been referred to the Committee on Interstate and Foreign Commerce, and specifically to the Subcommittee on Communications. The Bill must receive a hearing there before it can be sent to the House for action. If the Bill is to receive a hearing, however, the Congress must be made aware of our support for such legislation – support which you can demonstrate by writing to the Chairman of the Subcommittee on Communications, The Honorable Torbert H. Macdonald.*

The letters do not have to be long, although background information on your (or your neighbors') RFI problems could be important. Even a note to the effect that you support H.R. 7052 and *respectfully request an early hearing* would be a valuable contribution. It would also be a good idea to send a short letter to your own Congressman, indicating your desire that he support H.R. 7052. Remember that previously introduced RFI legislation never made it through Congress – don't let that happen this time.

Consumers are becoming increasingly aware of the RFI problem, so the time is right for legislation such as that proposed by Rep. Vanick. Amateurs have known for a long time that the majority of RFI problems are not due to interference *per se*, but are due to the interception of signals by devices which were not designed to operate in today's rf environment. The only way to eliminate 90% of the RFI problems is through legislation such as H.R. 7052 which could eventually require the manufacturers to correct those design deficiencies which lead to unnecessary interference.

Now is the time to lend your support to this vital effort. Write today, and make your voice heard.

Jim Fisk, W1DTY
editor-in-chief

*The Honorable Torbert H. Macdonald, Chairman, Subcommittee on Communications, Room R331, Rayburn House Office Building, U.S. House of Representatives, Washington, D.C. 20515.

presstop

FIRST WORKING GROUP MEETING in preparation for the 1979 World Administrative Radio Conference (WARC) went very well with nine "Task Forces" now set up and operating: ITU Rules and Regulations and Technical Criteria (W3FU), Military Liaison (W4FZ), Liaison with Other Services (W4ZC), Basis and Purpose (W6APW), 0-4 MHz Spectrum (W2QD), 4-27 MHz Spectrum (W3OKN), 27-1296 Spectrum (KH6IJ), and 1296 and Up Spectrum (Task Force Chairman shown in parenthesis).

Task Force Rosters were made up from those present, with additional membership still solicited from anyone willing and able to contribute time and expertise to this important effort. Contact Task Force leaders of the groups to which you wish to contribute. Task Force meetings will be held during the next few months, and a second meeting of the entire Working Group has been scheduled during the ARRL National Convention in Reston, Virginia in September.

Proposed New HF Amateur Bands got some encouragement during the general discussion, as preliminary views were expressed that the fixed services will likely be giving up enough frequencies in the HF range to accommodate the increased desires of international broadcasters plus at least some of the proposed new ham bands. However, the problem may very well not be finding the frequencies, but be one of marshalling sufficient support for the amateur service among the many new ITU members!

"DUAL LADDER" WAS REJECTED as ARRL Directors met to prepare League response to FCC's amateur restructuring docket. In a lengthy meeting in which the membership poll played a prominent part, the League leadership also came out against the loss of privileges proposed in Docket 20282 and recommended that present Conditionals and Technicians be permitted to renew their licenses indefinitely without reexamination. Although they supported the Communicator (they prefer "Basic" Amateur) class, they would include CW "recognition" in its requirements.

Member Survey Results, broken down in several ways, had been provided to the Directors for study well before the meeting. Of the approximately 56000 ARRL members who participated in the poll, about 20% also provided their Directors with written comments and few expressed any enthusiasm for either the dual ladder or loss of privileges.

AMATEUR DOCKETS starting to move at the FCC. Repeater Linking (Docket 20073) is reported on its way to the Commissioner's agenda, and Repeater Crossbanding (Docket 20113) is not far behind it. Action on Extra Class requesting specific callsigns (Docket 20092) is also expected soon.

Still Pending are RACES (19723), HIRAN (20147) and Repeater Automatic Control (20112). At least one new docket — "exotic emissions" — is in preparation.

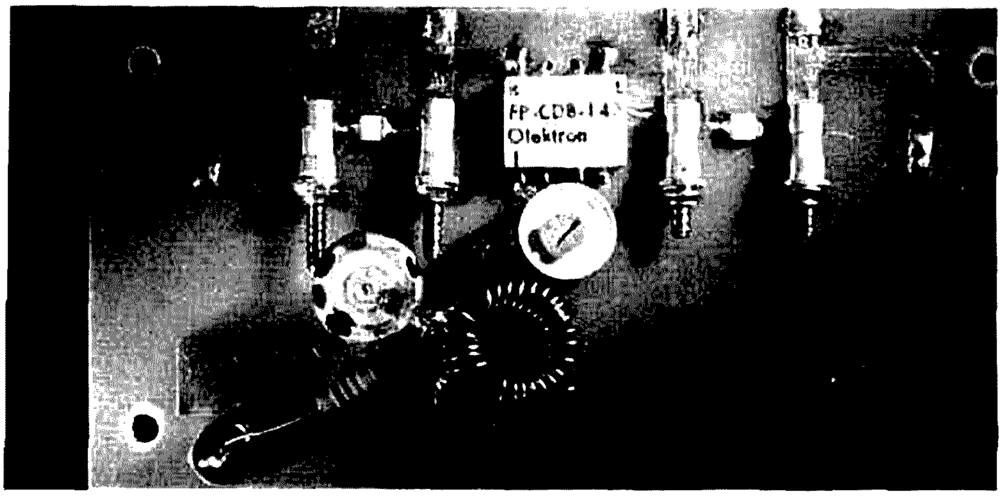
Docket Moving Unlicensed Hand-Helds from 27 MHz to just below six meters may be decided soon. Docket 20119 proposed putting the under-100 mW portables on discrete channels that alternate with already assigned MARS channels in the same range and as yet MARS authorities seem to have raised no objection.

FCC's Amateur and Citizens Division has been looking out for amateur interests, however — they've urged that superregen receivers be prohibited because of their potential interference to amateur operations on the low end of the 50-MHz band.

HAM RADIO, QST CHANGING FORMAT to 8½x11-inch page size effective with January, 1976 issues. Much justification for the change is economic — most periodicals, including practically all in the electronic field except for amateur radio, are in the larger format so printing, paper, handling, and the like should all be less expensive. Hopefully, the savings from these changes can delay future subscription rate increases.

OSCAR 7 is still suffering from occasional unexplained mode jumping. Any OSCAR user observing such a switch when it actually occurs should note the exact time and date and report them to AMSAT. Non-amateur satellites have had similar problems and AMSAT is attempting to correlate OSCAR 7's problem with theirs.

Russian Participation in OSCAR programs being discussed — two Russians, one a ham, spent two days with AMSAT representatives recently reviewing possible cooperative activities.



how to use double-balanced mixers on 1296 MHz

Modern commercial uhf double-balanced mixer modules improve the performance of 1296-MHz converters

Paul Shuch, WA6UAM, San Jose, California 95124

Diode balanced mixers have received considerable attention as bilateral mixers for transceive converter applications. In a previous article,[1] I outlined a transceive converter for 1296-MHz ssb including construction details for two suitable homebrew single balanced-mixers. Recent price breakthroughs have now brought within the reach of the serious experimenter several commercial double-balanced mixers which are adaptable for transmit, receive and transceive conversion well into the microwave region. This article describes the use of such mixers in the 23 cm band.

mixer modules

Since a number of manufacturers offer flatpack double-balanced mixers with identical lead arrangements, a circuit board can be designed which will accommodate a variety of mixer modules. These mixers, some of which are listed in table 1, vary primarily in conversion loss and power handling capabil-

ity. All of the mixers listed here will withstand the injection levels used in the 1296-MHz ssb transceive converter (i.e., 40 mW of local-oscillator injection and 12 mW PEP of applied i-f power in the transmit mode).

functionally equivalent to a popular design which many amateurs have built for lower frequency applications.[2] In theory, all that is necessary to use the mixers is to mount the flatpack device on a circuit board containing connectors

table 1. Partial listing of flatpack double-balanced mixers suitable for use on the 1296-MHz band. Conversion loss shown is worst cast for the specified frequency range; it may be less at spot frequencies within the overall range. Mixers are listed in order of ascending single-quantity price. Manufacturer's addresses are listed below.

manufacturer	model	frequency range	conversion loss	approximate price
Olektron Corporation	FP-CDB-145	0.5-1350 MHz	9.0 dB	$29
Merrimac Industries	DMF-2A-750	50-1500 MHz	9.0 dB	$40
Vari-L Company	DBM-158	500-1500 MHz	7.5 dB	$50
Lorch Electronics	FC-200ZF15	10-1500 MHz	8.5 dB	$59
Watkins-Johnson	M4A	10-1500 MHz	8.5 dB	$60
Anzac Electronics	MD-614	600-2000 MHz	7.5 dB	$75

Anzac Electronics, 39 Green Street, Waltham, Massachusetts 02154
Lorch Electronics, 105 Cedar Lane, Englewood, New Jersey 07631
Merrimac Industries, 41 Fairfield Place, W. Caldwell, New Jersey 07006
Olektron Corp., 6 Chase Avenue, Dudley, Massachusetts 01570
Vari-L Company, 3883 Monaco Parkway, Denver, Colorado 80207
Watkins-Johnson Company, 3333 Hillview Avenue, Palo Alto, California 94304

The commercial flatpack double-balanced mixers all contain a ring of four hot-carrier diodes, typically in a beam-lead pill package, along with two wideband balun transformers. They are

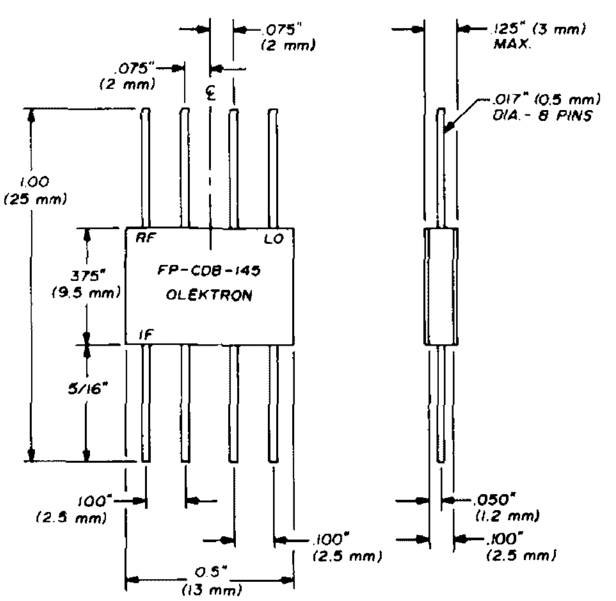


fig. 1. Physical dimensions and pin layout of typical flatpack double-balanced mixer module.

(to each of the three ports) and provide a good ground path for those terminals which must be grounded. Other features can be added to the circuit board to enhance its performance in a particular application.

The circuit board presented in this article is designed to permit transceive conversion to the 23 cm band while providing image filtering at the rf port, spurious response filtering at the local-oscillator port, and resonating the i-f port to 28 MHz.* Provisions are also made for monitoring mixer current in the dc return to the i-f port.

The prototype of this assembly uses an Olektron model FP-CDB-145 flatpack mixer. Its physical configuration (which is typical of all the flatpack mixer modules) is shown in fig. 1. The Olektron device is the least expensive of the

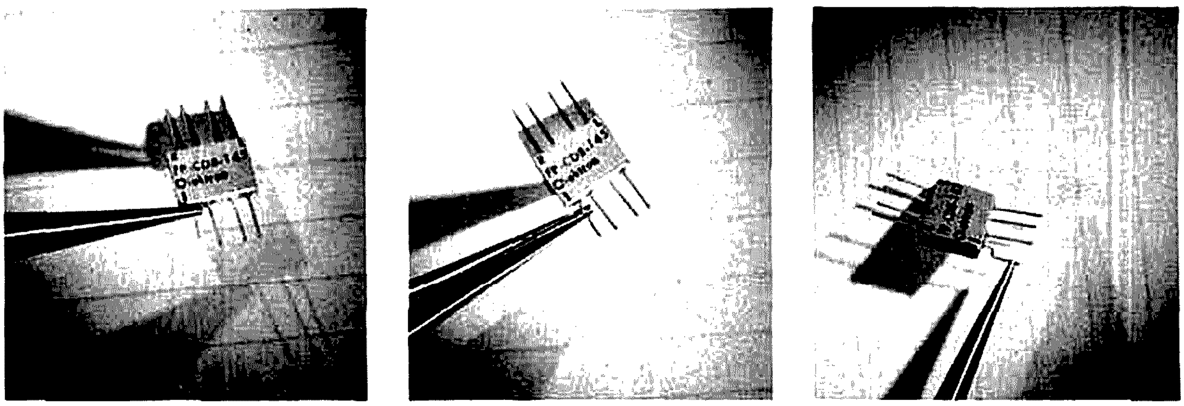
*Other intermediate frequencies may be used by suitable modification of components L5, L6 and C7 in fig. 2.

mixers listed in table 1, and offers electrical characteristics which are completely satisfactory for the 1296-MHz transceive application (see table 2). Should lower conversion loss be required, you may wish to substitute one of the more expensive mixers listed in table 1. However, as you might expect, each 1 dB of improvement in conversion loss is offset by a corresponding 1 dB increase in cost.

bandwidth considerations

When applying any of the readily available commercial mixers to narrowband service, you should be aware of their inherently broad frequency response. The Olektron unit, for example, is designed for operation from 500 kHz to 1350 MHz. The very bandwidth which is so beneficial in many applications may well prove a detriment here as an absence of selectivity at the rf and local-oscillator ports invites out-of-band spurious responses. My homebrew mixer designs have incorporated frequency-selective circuitry at the various ports. To alleviate interference problems related to unnecessary bandwidth, such filtering should be added externally to any of the commercial mixers. Unless the mixer module is already provided with connectors, this filtering is readily incorporated on the circuit board on which the mixer is mounted. In fig. 2, components L1 through L6 and C1 through C7 serve this purpose.

As was mentioned previously, it is possible to use any of the flatpack mixers simply by mounting them on a board containing the appropriate connectors for interfacing to the three ports. Upon attempting this simplistic approach in a 1296-MHz transceive converter, however, I found myself with more birdies than the Audubon Society, and enough image to run for office. Only upon incorporating the filtering provisions of fig. 2 did the melange of signals emanating from the mixer become manageable.



To bend the leads on the delicate double-balanced mixer package, first grasp the pins with long-nose pliers very close to the package and bend the lead downward at 90 degrees (left). Then grasp the lead at the bend and form the lead straight out from the package (center). The desired lead configuration is shown on the right.

diode current monitoring

Often it is desirable to monitor mixer diode current during converter operation. This is especially useful while tuning local-oscillator chains, or for determining the adequacy of the i-f injection applied in the transmit mode. Unfortunately, most commercial mixers make no direct provision for monitoring diode current. Again, this feature may be added to the board on which the mixer is mounted.

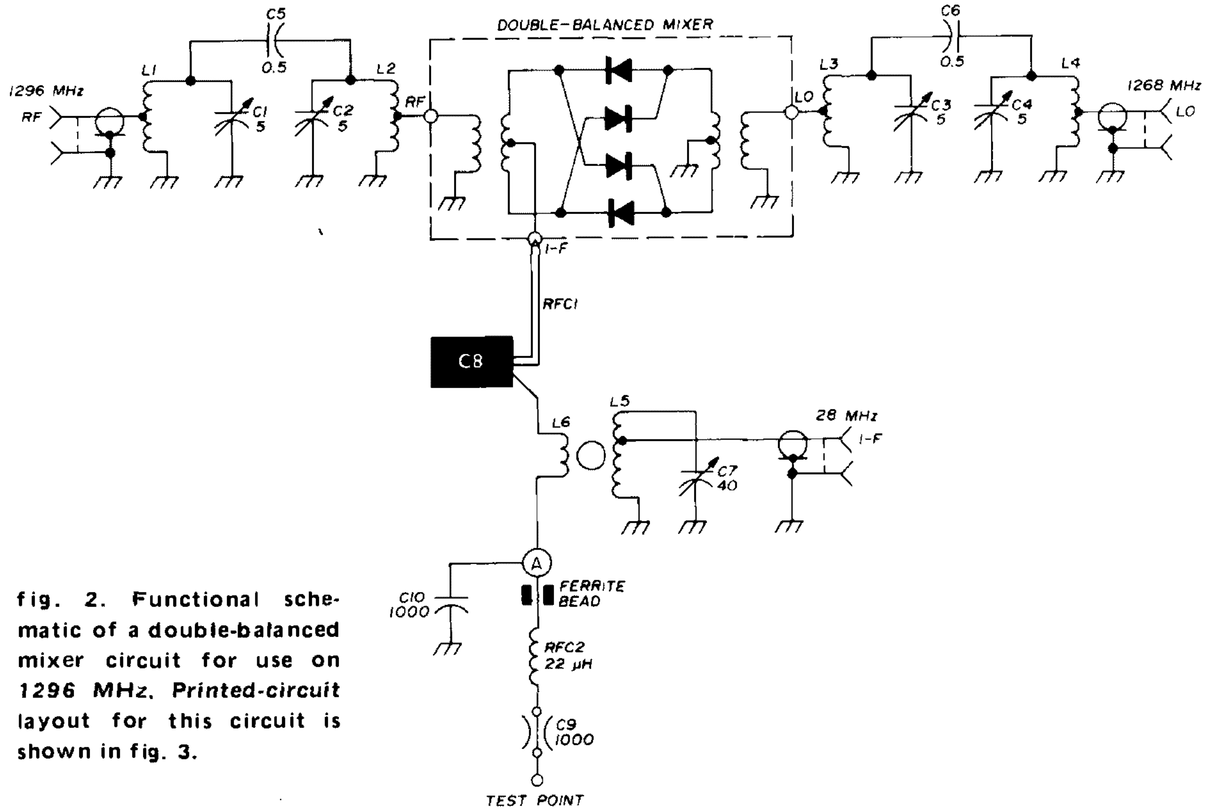


fig. 2. Functional schematic of a double-balanced mixer circuit for use on *1296 MHz. Printed-circuit* layout for this circuit is shown in fig. 3.

C1-C4	1-5 pF ceramic piston trimmer (Triko 201-01M or equivalent)
C5,C6	0.5 pF chip capacitor (ATC-100 or equivalent)
C7	10-40 pF ceramic trimmer
C8	25-ohm quarter-wavelength open-circuited microstripline, 0.3" (7.5mm) wide, 1.14" (29mm) long
C9	1000 pF feedthrough capacitor
C10	1000 pF standoff capacitor
DBM	double-balanced mixer module (Olektron FP-CDB-145 or similar, see table 1)
L1-L4	44-ohm 0.1 wavelength shorted microstripline, 0.125" (3mm) wide, 0.5" (13mm) long, tapped 0.2" (5mm) from grounded end
L5	1.2 µH. 19 turns no. 22 AWG on Amidon T50-10 toroid core, tapped 5 turns from cold end
L6	5 turns no. 22 AWG on same core as L6
RFC1	50-ohm microstripline, 0.1" (2.5mm) wide, 1.20" (30.5mm) long
RFC2	22 µH molded inductor

The circuit of **fig. 2** incorporates a current monitoring provision. Point A is normally connected directly to ground. If the ground is omitted, rectification by the diode quad results in a dc component; relative current may be monitored by measuring the voltage drop with a sensitive vtvm connected between point A and ground. Additionally, proper operation of the mixer requires point A to be at rf ground. Therefore, effective bypassing of *all* rf components must be provided. Bear in mind that the local-oscillator to i-f isolation is only on the order of 16 dB. Thus, with 40 mW of local-oscillator injection, and without adequate bypassing, a disruptive 1 mW of local-oscillator energy will appear on the bias test point.

In my application the local-oscillator and rf frequencies are sufficiently close (1268 and 1296 MHz, respectively) to be adequately bypassed by capacitor C8, an open-circuited quarter-wavelength microstripline of low characteristic impedance. Grounding the i-f component (28 MHz in this case) is accomplished with bypass capacitor C10. In

addition to the bypassing components, an rf choke and ferrite bead are used to isolate the bias test point from any remaining i-f, local-oscillator or rf signals.

tors of the BNC, TNC or SMA variety. **Fig. 6** details the fabrication of these launchers and shows a method for mounting them to the circuit board.

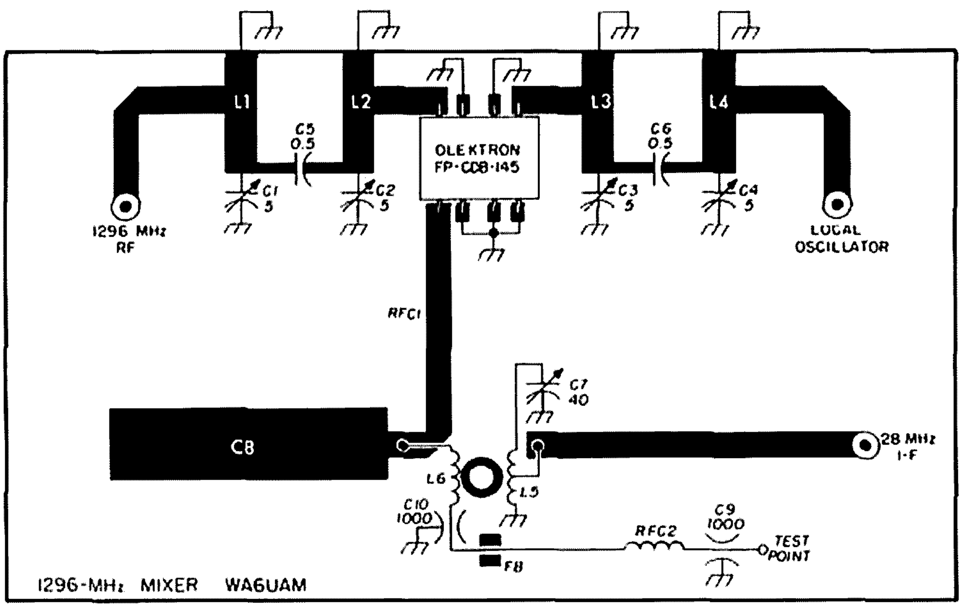


fig. 3. Double-balanced mixer circuit for 1296 MHz which uses microstripline construction. Component details are listed under fig. 2. Full-size printed-circuit board is shown in fig. 4.

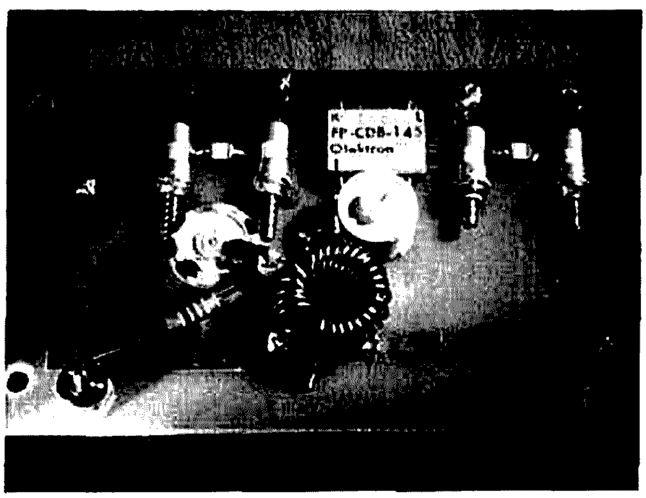
construction

Construction of the printed-circuit mixer assembly is shown in **fig. 3**. Full-sized artwork for the circuit board is shown in **fig.** 4. The microstripline dimensions are for use with 1/16 inch (1.5mm) thick G-10 fiberglass-epoxy board, double clad with 1 ounce copper. When building the board be sure to leave an unetched ground plane on the opposite side of the board.

Microstripline inductors L1, L2, L3 and L4 are grounded by pieces of copper foil which are wrapped around the board edge to the ground plane as shown in **fig.** 5. Holes are drilled in the board so the tuning-screw terminals of capacitors C1, C2, C3 and C4 can be connected directly to the ground plane. Minimum lead lengths are imperative.

Interfacing of the rf, i-f and local-oscillator ports is accomplished by microstripline launchers which can be built from flange-type coaxial connec-

Mounting of the fragile double-balanced mixer module should be deferred until all other components are in place. Holes are then drilled to permit direct through-the-board *grounding* of all mixer pins not connected to micro-



Double-balanced mixer layout for 1296 MHz, showing component placement on the printed-circuit board (see fig. 3).

table 2. Electrical specifications for the Olektron model FP-CDB-145 double-balanced mixer module.

Frequency response		RF port	0.5 to 1350 MHz	
		LO port	0.5 to 1350 MHz	
Conversion loss, typical	500 MHz	6.5 dB		
	1000 MHz	7.5 dB		
	0.5 to 1350 MHz	9.0 dB		
		LO/RF	RF/IF	LO/IF
Isolation, minimum	500 MHz	30 dB	20 dB	20 dB
	1000 MHz	20 dB	15 dB	15 dB
Local-oscillator power		+7 to +13 dBm (5 to 20 mW)		
Diodes used		hot carrier		
Temperature		−54°C to +100°C		

striplines. The leads to the rf, local-oscillator and i-f terminals are bent down at right angles to the plane of the substrate, then out at 90 degrees so they lie on top of the microstriplines to which they are soldered.

When you're bending the leads of the mixer module, use extreme care so you do not fracture the delicate metal-glass lead seal. It is advisable to grip each lead with a pair of small needle-nosed pliers at the point where the lead just exits the flatpack; then bend the lead down on the far side of the pliers. For the local-oscillator, rf and i-f leads use the pliers to grip each lead just below the first bend when making the second bend. See the accompanying photograph for clarification (page 10).

tuneup and operation

In addition to bandpass filtering, proper adjustment of the resonators at each port of the double-balanced mixer assembly will provide the required impedance matching. An excellent technical note from Anzac Electronics[3] describes the disastrous side effects of improperly terminating the various ports of a double-balanced mixer. As can be seen from **table 3**, a worst-case combination of reactive mismatches to all three ports can result in an overall degradation in mixer conversion loss of 3.5 dB, with a corresponding increase in third-order intermodulation products up to 30 dB.

When tuning the mixer assembly it is imperative to adjust each tuned circuit so that its associated mixer port sees a nonreactive 50-ohm termination. This is accomplished by adjusting the local-oscillator filter for maximum diode current, and adjusting the rf and i-f filters for minimum single-sideband conversion loss.

Initial adjustment of the local-oscillator port is most easily accomplished

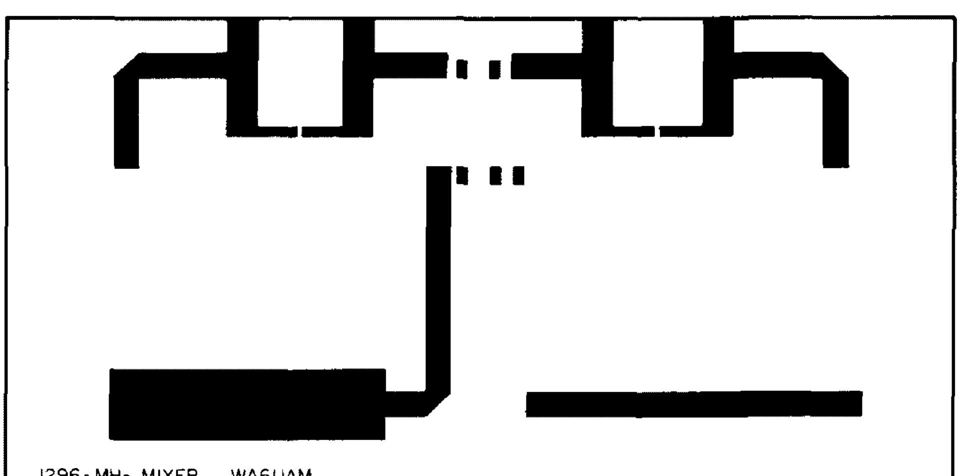


fig. 4. Full-size printed circuit board for the 1296-MHz double-balanced mixer. Material is 1/16" (1.5mm) G-10 fiberglass-epoxy board, double clad with 1 ounce copper.

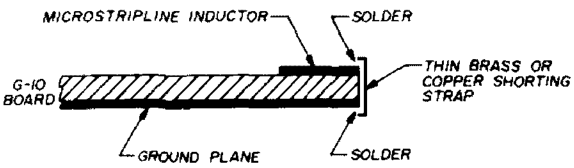


fig. 5. Method of properly grounding inductors L1, L2, L3 and L4 with thin shorting straps to the ground plane on the reverse side of the board.

by coupling in a +5 to +13 dBm local-oscillator signal (3 to 20 mW) and adjusting trimmers C1 and C2 alternately for maximum relative diode current, as indicated on a sensitive vtvm at the bias test point. Preliminary adjustments of the rf filter, to assure that the mixer is not tuned up on the image frequency, should be performed by coupling 10 milliwatts or so of 1296-MHz energy from a signal generator into the rf port and adjusting C3 and C4 for a maximum reading at the test point with *no* local-oscillator injection. If a grid-dip oscillator, tuned to 28 MHz, is link-coupled through a coax cable to the mixer's i-f connector, the i-f tank circuit can be tuned for a dip with C7.

Final adjustments to C3, C4 and C7 can be made by connecting the mixer to a local-oscillator chain, i-f receiver and weak signal source, and tweaking for optimum signal-to-noise ratio. *Do not* adjust C1 and C2 at this time as optimizing diode current with the local oscillator connected has the effect of impedance-matching the local-oscillator port. While it is true that the mixer's conversion loss will vary with local-oscillator injection level (see fig. 7), to minimize third-order intermodulation products, changes in the local-oscillator injection level should be accomplished by padding the output of the oscillator, *not* by mismatching the mixer's local-oscillator port.

table 3. Effect of reactive terminations on double-balanced mixer performance. Mixer terms are defined below. (Data courtesy the Engineering Department, Anzac Electronics.)

termination condition	conversion loss	rf compression level	rf de-sensitization level	harmonic modulation products	third-order IM products
RF Port = 50 ohms IF Port = reactive load LO Port = 50 ohms	can vary ±3 dB	can vary ±3 dB	can vary ±3 dB	can vary ±20 dB	can vary ±20 dB
RF Port = 50 ohms IF Port = 50 ohms LO Port = reactive source	no effect if LO drive adequate	no effect if LO drive adequate	no effect if LO drive adequate	can vary ±10 dB	can vary ±10 dB
RF Port = reactive source IF Port = 50 ohms LO port = 50 ohms	typically ±0.5 dB for 2:1 vswr*	±0.5 dB	±0.5 dB	no first order effect	no first order effect

*Increases proportional to mismatch.

Harmonic modulation products: Output responses caused by harmonics of the local-oscillator and rf signal and their mixing products.

RF compression level: The rf input power level that causes the conversion loss to increase by 1 dB.

RF desensitization level: The rf input power of an interfering signal that causes the small-signal conversion loss to increase by 1 dB.

Intermodulation products: Harmonically related distortion products caused by multiple rf signals and their harmonics mixing with each other and the LO producing signals at new frequencies.

The most elegant method of adjusting this mixer, of course, is to tune all ports for minimum indicated noise figure on an automatic noise meter system. However, those amateurs blessed with access to such equipment are cautioned to verify their results with a crystal-controlled weak-signal source to guard against resonating the rf port's filter at the image frequency. When properly adjusted, this mixer's image rejection is on the order of 18 dB.

My prototype mixer assembly, shown in the photographs, indicates a ssb noise figure of 10.5 dB, measured with a Hewlett-Packard 340 B Automatic Noise Meter and an AIL 7010 argon discharge noise head. Local-oscillator injection at optimum noise figure is +10.5 dBm. The measured noise figure of the dual-gate mosfet i-f amplifier which followed this mixer during noise measurements is 1.0 dB. Therefore, if you accept the published figure of 9.0 dB for the conversion loss of the flatpack mixer module, the filter loss at the rf port is on the order of 0.5 dB.

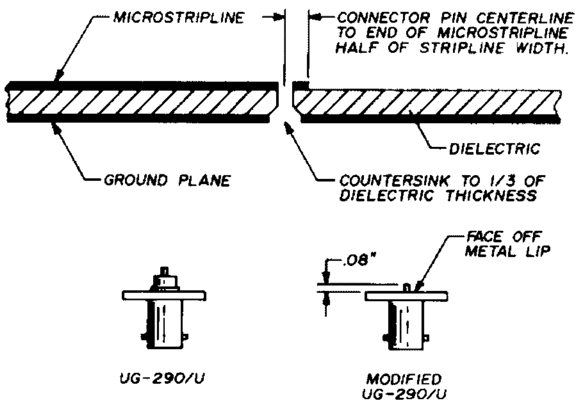


fig. 6. Technique for modifying flange-type coaxial connectors for microstripline launchers. Same method can be used with BNC, TNC or SMA coaxial connectors.

The 9.5 dB conversion loss of this mixer assembly is on a par with my homemade mixer designs while offering the improved isolation and reduced harmonic modulation product density characteristic of the double-balanced arrangement. Performance of this mixer in the receive mode is wholly satisfactory for line-of-sight communications on the 23 cm band. Should beyond-the-horizon communications be desired, a single stage of low-cost preamplification[4] will reduce overall system noise figure below 5 dB. With the addition of a low-noise stage I have brought my system noise figure down to 2.2 dB.

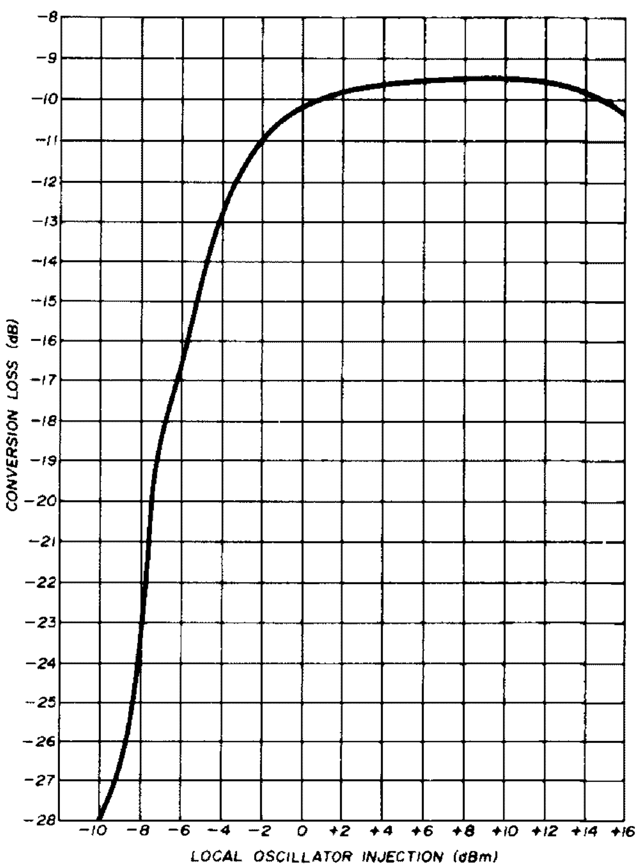


fig. 7. Measured conversion loss of the 1296-MHz double-balanced mixer circuit as a function of local-oscillator injection level.

references

1. H. Paul Shuch, WA6UAM, "1296 MHz SSB Transceiver," *ham radio*, September, 1974, page 8.
2. William J. Ress, WA6NCT, "Broadband Double-Balanced Modulator," *ham radio*, March, 1970, page 8.
3. "Reactive Loads — The Big Mixer Menace," Technical note, Anzac Electronics, 39 Green Street, Waltham, Massachusetts 02154.
4. H. Paul Shuch, WA6UAM, "Microstripline Preamps for 1296 MHz," *ham radio*, April, 1975, page 12.

ham radio

universal tone encoder

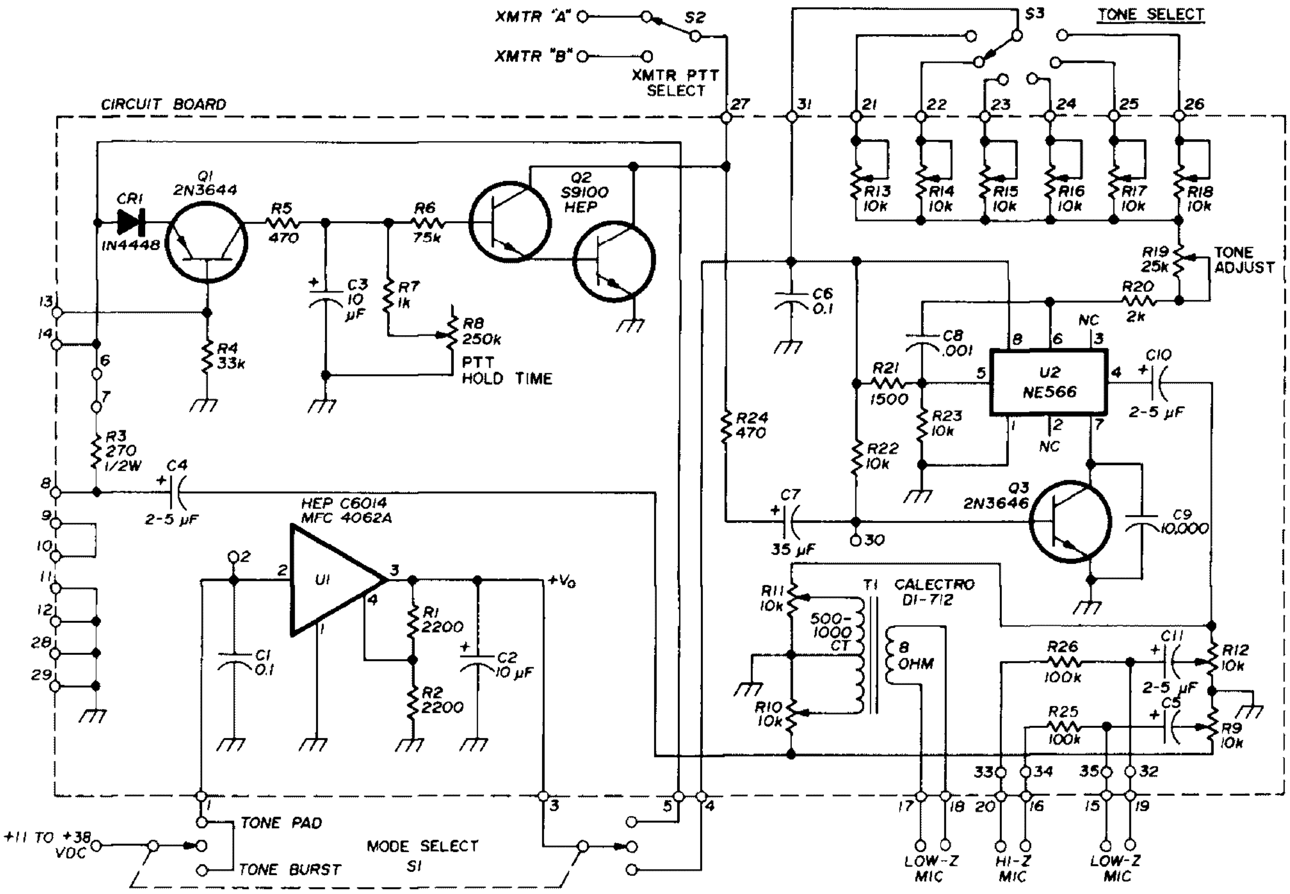
for vhf fm

A straight-forward tone encoder which provides both Touch-Tone signaling and tone-burst operation

Larry McDavid, W6FUB, Anaheim, California

The availability of low cost Touch-Tone* decoders, autopatch and tone-burst-operated repeaters has resulted in renewed interest in tone encoders. However, much confusion exists as to how tone pads and burst oscillators can be built and interfaced with vhf fm transceivers. The unit described here provides both Touch-Tone and tone burst, yet is stable and easily constructed. Three types of audio outputs are provided as well as automatic PTT hold during Touch-Tone dialing. The unit can be operated from internal batteries or from automotive or other 12-volt systems. All components are commonly available and construction is simplified by the use of a printed-circuit board.

*Touch-Tone is a registered trademark of the Western Electric Company.



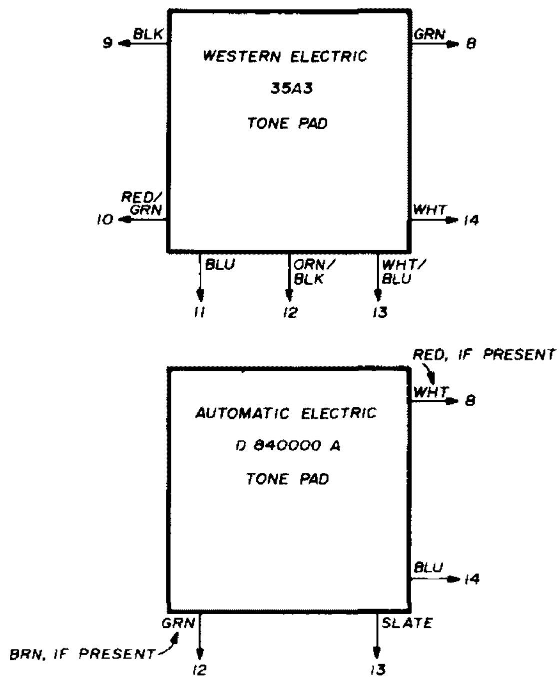


fig. 1. Circuit of the universal tone encoder. This circuit provides both high- and low-impedance microphone outputs, and may be used with more than one transmitter. Before using either of the tone pads, remove the 5100-ohm resistor (see text). Resistor R1 is selected to set the regulated voltage, V_O, at +8.5 volts. Resistors are 1/4 watt unless otherwise noted.

output coupling

Two sets of independent level controls allow operation with two different transceivers without readjustment. One high-impedance and two types of low-impedance outputs are provided. Independent Touch-Tone and tone-burst outputs may be combined by jumpering terminals 33 to 34 and 32 to 35 (see **fig. 1**).

High-impedance inputs may be driven by connecting terminal 16 or 20 directly to the microphone line. Resistors R9 and R12 adjust the output levels for Touch-Tone and tone burst, respectively. If the drive level is too low the value of R25 or R26 may be reduced.

Low-impedance inputs may be driven by connecting terminals 17 and 18 in *series* with the microphone line. Resistors R10 and R11 adjust the output levels; no change in microphone audio quality or deviation will result. This has

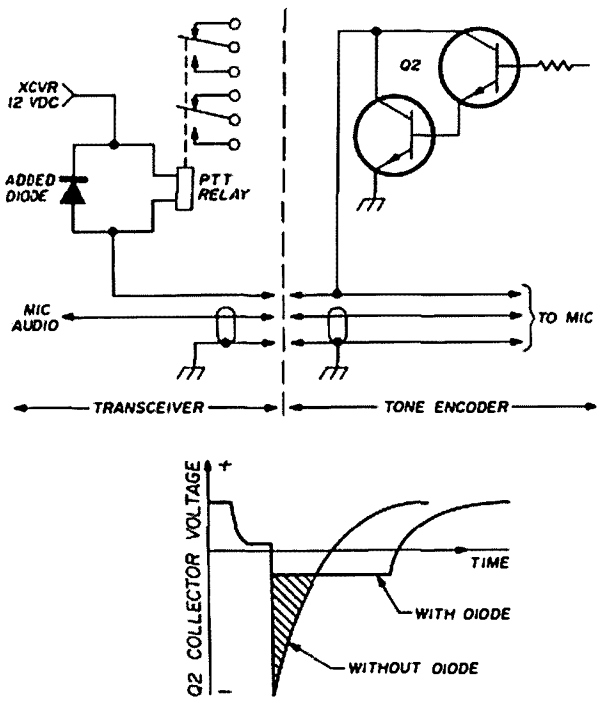


fig. 2. Correct wiring for the PTT relay in your fm transceiver. Proper waveform on the collector of transistor Q2 is shown below the schematic.

an advantage in many transceivers: a microphone shorted by the PTT switch will not pick up background noise during Touch-Tone signaling. A shorting type microphone is not required for operation, however. The microphone may be plugged into the tone encoder enclosure and a shielded cable run to the transceiver. Another low-impedance output is available at terminal 15 or 19 if transformer coupling is not desired. This output should be connected directly to the microphone line. Unused outputs may have the associated components omitted.

tone pad encoder

Wiring for Western Electric and Automatic Electric tone pads is shown in fig. 1. One modification is required to the pad: the removal of the 5100-ohm resistor used for reducing the local handset tone signaling volume. One end may be clipped or the resistor may be entirely removed. This provides a pair of normally-shorted contacts that are open when any button is depressed.

Some Automatic Electric tone pads have an internal diode bridge to allow

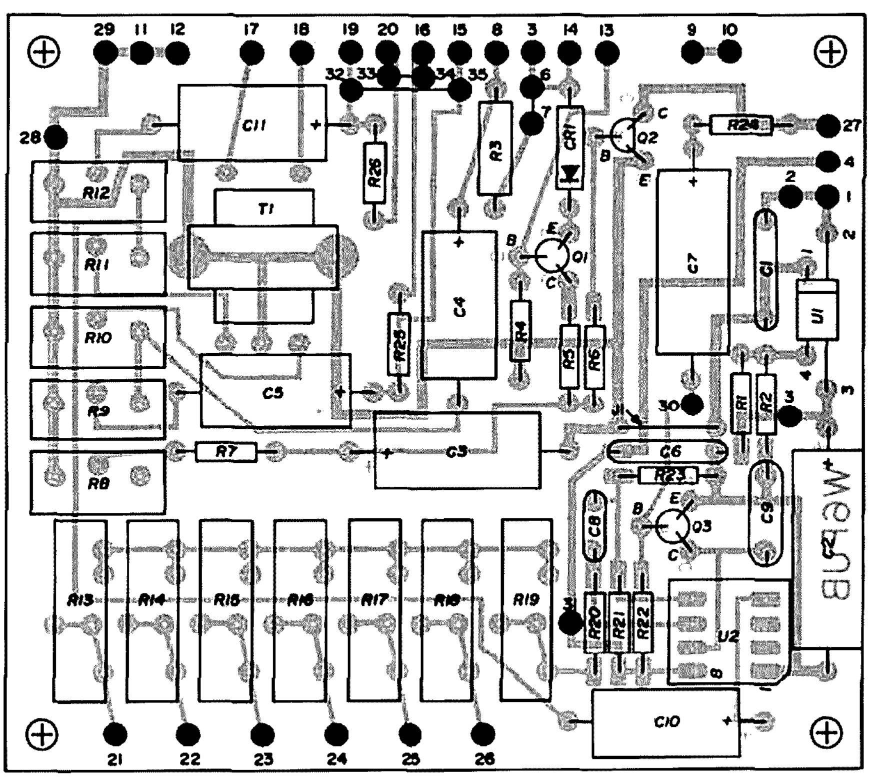
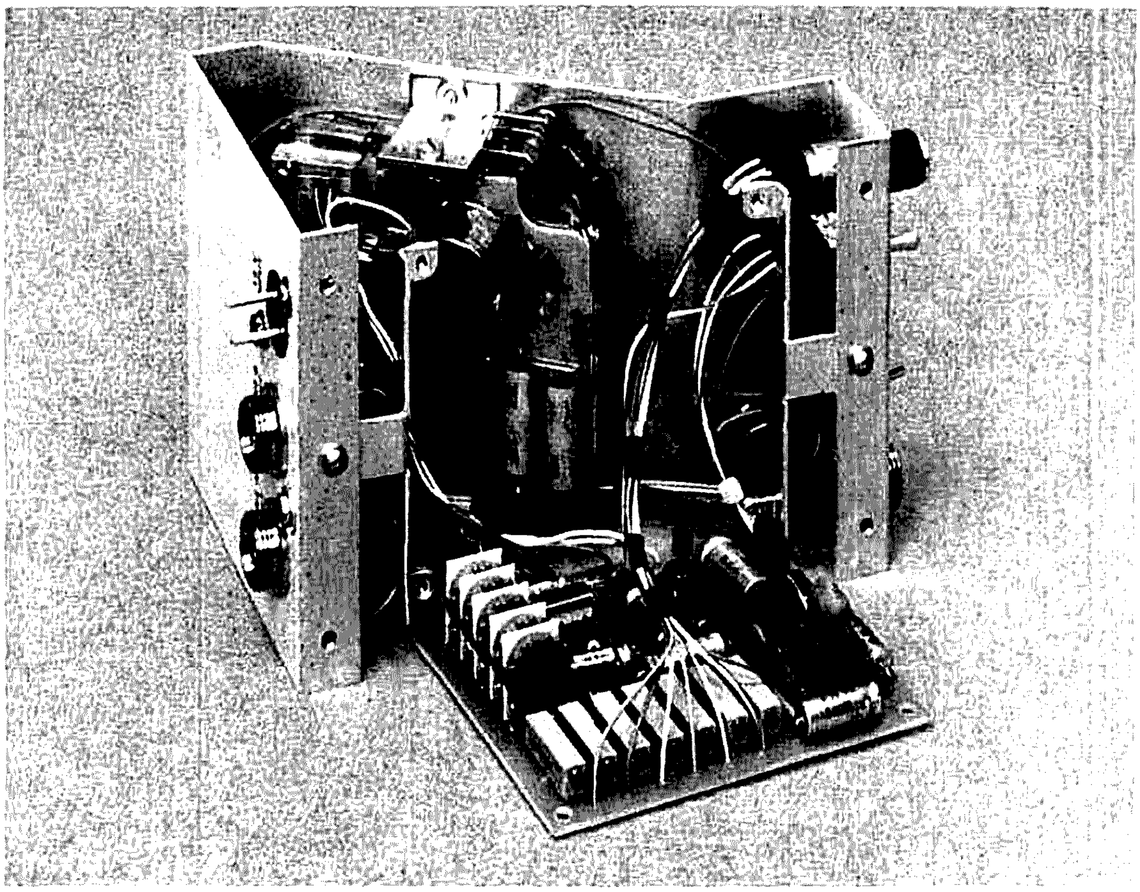


fig. 3. Component layout for the tone encoder circuit board. Full-size printed-circuit board is shown in fig. 4.

operation with either polarity input voltage. These wires, red and brown, may be used rather than the green and white. Any unused wires from either pad should be cut off or taped to prevent shorting.

of C3 will yield longer delays. This approach is not new and has been tested for some time with no problems.[1] The delay interval should be set just long enough to dial a tone sequence without rushing.



Construction of the universal tone encoder. The printed-circuit board is shown in the foreground.

Transistor Q1 is normally biased off by CR1, and the regulated supply voltage, V_o, is applied to the base through terminals 13 and 14. When any tone button is depressed, terminal 13 is opened, turning on Q1 and Q2 and charging C3. The PTT relay is operated by ground return through transistor Q2.

When the tone button is released, Q1 is again turned off. Transistor Q2, however, is held on until the charge on C3 is bled off through R6, R7 and R8. The PTT hold time is adjusted by R8. Delays up to two seconds are possible with the components shown. Increasing the value

Operation of this circuit assumes that the transceiver PTT relay is wired as shown in fig. 2. The collapsing field of the relay coil will cause the voltage spike on the PTT line when the relay is opened. Although no trouble has been experienced by several users, damage to Q2 could result. The spike can be eliminated by installing a diode across the relay coil; although relay drop-out time will increase, it should not be noticeable. Maximum relay current must be less than 300 mA.

Transceivers which use diode PTT switching will generally work with this

circuit without change. If any switching difficulty is experienced, a small reed relay could be used to switch the transceiver; Q2 would operate this relay coil.

Switch S2 allows the encoder to be used with two transceivers by selecting the appropriate PTT line. CR1 may be any silicon diode; Q2 should not be substituted. The other components are not critical.

R19 for approximately 2500 Hz, using a counter connected at the junction of C10 and R11. Then adjust R13 through R18 for the desired burst frequencies selected by switch S3. Tone-burst duration is about 0.4 second, but may be varied by changing the value of C7. The oscillator has been used for MCW by connecting a key between terminal 30 and ground.

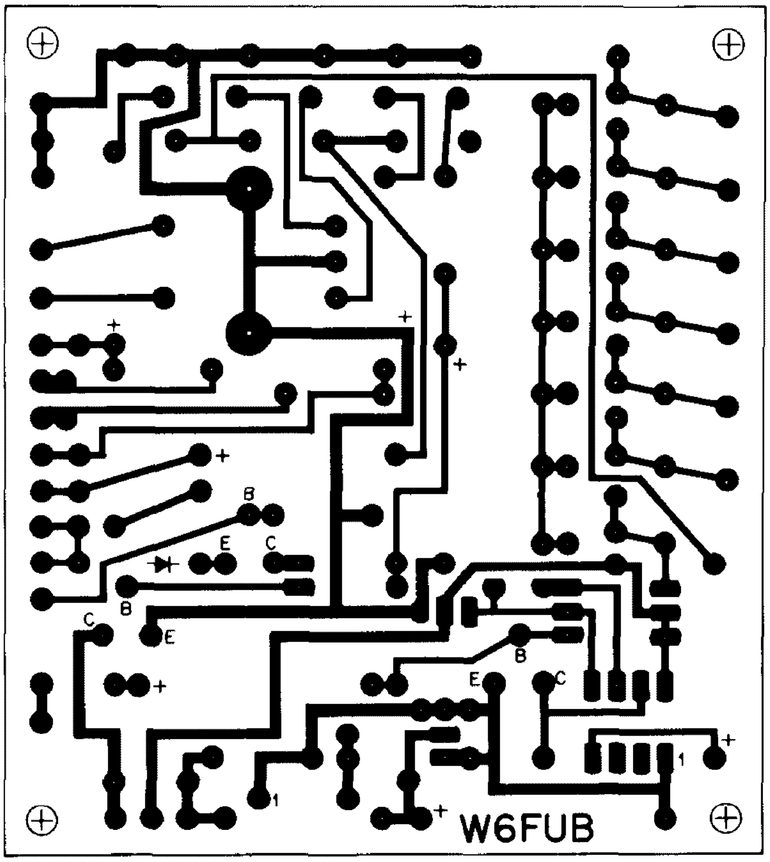


fig. 4. Full-size printed-circuit board for the universal tone encoder. Component layout is shown in fig. 3.

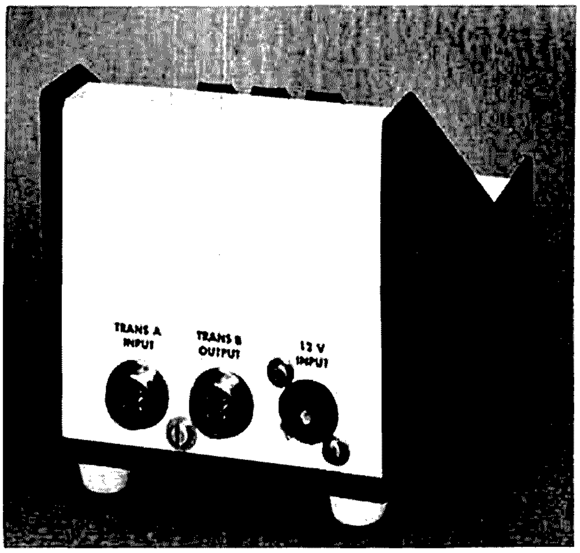
tone-burst oscillator

This circuit was first described elsewhere and has proven stable and reliable.[2] A polystyrene capacitor was selected for C9; a mylar or disc may be substituted with some loss of temperature stability of burst tone frequency. Resistor R20 prevents accidental damage to U2 if all the tone adjust pots are set to minimum resistance.

The oscillator frequency is set by grounding terminal 30. Set R13 through R18 to minimum resistance and adjust

The burst oscillator, in particular, requires a well regulated power supply. To minimize the current drain, a low-cost IC regulator, U1, was selected. Input voltage variations do not cause excessive current drain as with a zener regulator. Inputs between 11 and 38 volts are satisfactory, allowing operation with either an auto battery or two internal 9-volt batteries in series.

Total current required is 11 mA for tone burst and 20 mA during tone pad signaling. A further advantage of the IC regulator is the elimination of generator



Rear view of the universal tone decoder showing the input and output connectors.

or alternator noise on the encoder output. Switch S1, a center-off dpdt, selects the mode of operation. Resistor R1 may be selected if required to set the regulated voltage V_0 to 8.5 volts.

construction

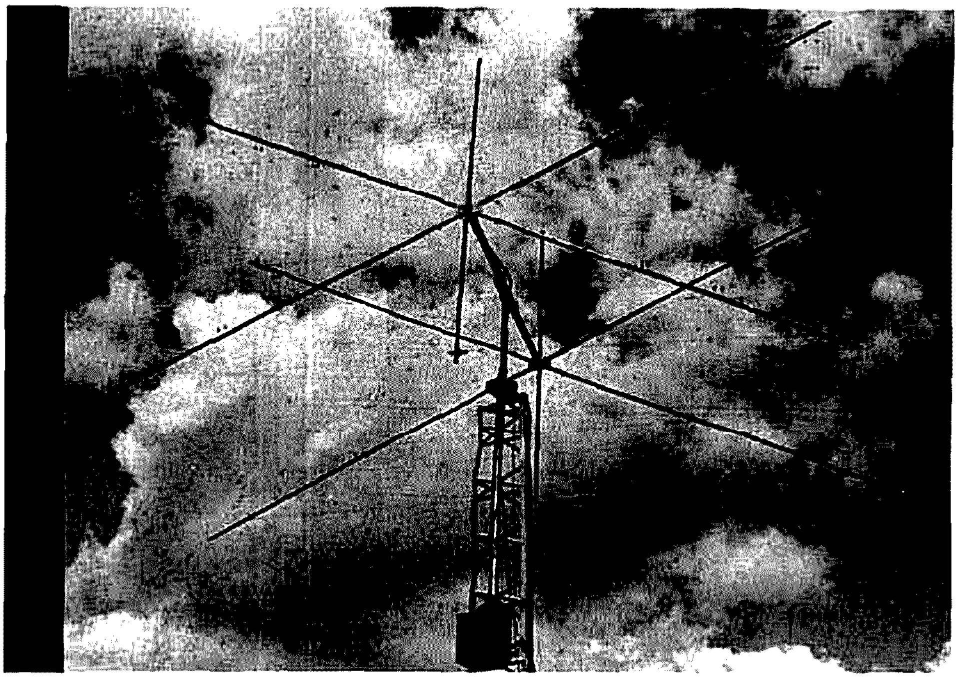
The internal packaging of the encoder is shown in the photographs. A LMB-007-446 cabinet was used. All components except switches and connectors mount on the 3.1x3.5-inch (79x89mm) printed-circuit board.* Fifteen-turn trimpots are shown for the tone-burst frequency adjustments. However, the board will also accept single-turn pots. A socket is used for U2; all other components are soldered in place. If only internal batteries are used, the 12-volt input connector on the rear panel may be omitted.

references

1. Roy Hejhall, K7QWR, "Solid-State Mobile Touch-Tone Circuit," *ham radio*, March, 1973, page 50.
2. Glen Dickson, WB5BAF, "A Tone-Burst Generator for Repeater Access, *QST*, April, 1974, page 30.

ham radio

*Drilled and plated printed-circuit boards are available from the author for $5.50, postpaid.



low profile three-band quad

The low profile quad offers several mechanical advantages over the usual three-band design

John P. Tyskewicz, W1HXU, Hartford, Connecticut 06112

Multiband coverage with a compact beam antenna is a necessity with most amateur operators and complements the modern station contained in a single cubic foot of space. After having achieved excellent quad performance with the LPQ 20-meter monobander,[1] an excursion to the 10 and 15 meter arena with this same type of antenna seemed mandatory. At the same time, some structural and electrical improvements were to be implemented.

The casualty rate among quads is rather high, and although my bamboo spreaders survived last winter's ice storm, it probably would have finished off a tri-bander with its additional surface area. Static load tests revealed the quad's inherent weakness when built with small diameter spreaders. The easiest solution seemed to be the addition of a vertical king post to each spider; the horizontal wires of each loop could then be attached to them to provide some useful load bearing. To accomplish this, a 1-foot (30.5cm) length of 1x1 inch (25x25mm) angle iron was bolted to the outer side of each spider and a wood king post 1-1/8 inch (28mm) OD by 9 feet, 4 inches (2.84m) long was installed. Number-8 (4mm) panhead sheet metal screws at appropriate points along the king post provided anchor points for the wire.

An aluminum or metal king post would provide even greater rigidity.

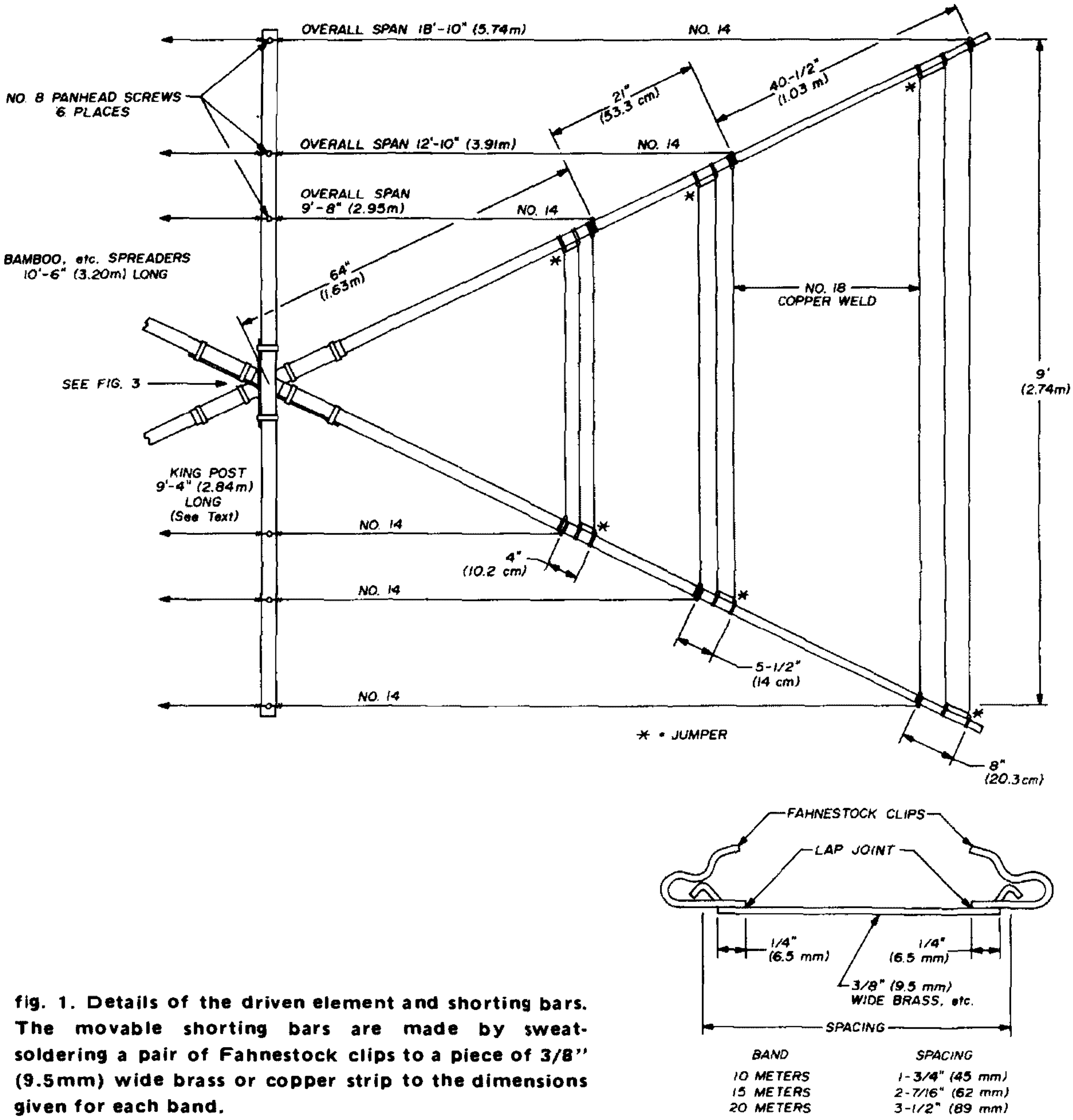


fig. 1. Details of the driven element and shorting bars. The movable shorting bars are made by sweat-soldering a pair of Fahnestock clips to a piece of 3/8" (9.5mm) wide brass or copper strip to the dimensions given for each band.

However, its presence might disturb the antenna electrically. To test its effect, the antenna was first built with the wooden king post described. After the tri-bander was completed and tuned to my satisfaction, a number-18 (1mm) wire was run the full length of the reflector king post and connected to all horizontal wires at their center (high current) points and to the spider and boom. This grounding did not upset the swr or field strength readings, so the driven element was also grounded in this same manner and with equally good results.

As an added benefit, static charge collection is reduced; now during a passing thunderstorm I am not disturbed by any snap-cracking within the confines of the tuner box! From these observations you can assume that a metal tubing king post could be used for better grounding and rigidity without affecting performance.

construction

To LPQ newcomers the spider shown in **fig. 3** is easier to make than the original type. If necessary, the parts can be assembled with 8-32 (M4) hardware

and taken to a welding shop for finishing. Close attention to dimensions will take care of the necessary 54 degree spreader angle. To attain good radiator-reflector alignment once the individual spiders are assembled, the following procedure is recommended: First, insert one spider (without spreaders) in the boom and drill through both boom and spider with a 1/8 inch (3mm) drill. Then follow with a number-7 (5.1mm) drill, and finish by tapping both pieces together with a ¼-20 (M7) tap. Fasten them with a machine screw, and continue to drill-tap the other holes.

When one spider is completed, insert the opposite spider into the boom and lay the entire assembly on a flat surface. Check the top of the spider arms with a level, and repeat the drill-tap operation. Before dismantling, index and identify the spider-boom positions with a center punch or small chisel marks.

The spreader clamps, **fig. 3**, are made from 20 gauge (1mm thick) stainless or galvanized iron stock, and they should be formed so as to leave a gap of approximately 1/8 inch (3mm) when fully tightened. The bamboo spreaders should be weatherized by spiral wrapping them with PVC electrical tape. I also wrapped one set of spreaders with 3/4 inch (19mm) wide paper masking tape, wiped them off with naptha, and coated them with latex paint. To date the paper covering has not deteriorated, is cheaper, and can color match the sky or your house.

elements

Fig. 1 shows in detail the driven element and its shorting bars. In building this "monster," you're confronted with the unpleasant task of locating and providing a large number of anchor points along the spreaders. Begin by matching up the bamboo poles and selecting the stronger pair for the upper set. Then you can either fasten a spreader to the proper spider arm and measure off the given radii, or measure from the butt end and allow for the difference. Wrap three layers of PVC tape at all anchor points.

The wire-loop anchors are formed as shown in **fig.** 2. The approximately 1/8 inch (3mm) ID two-turn loop is made by wrapping the number-16 (1.3mm diameter) galvanized iron wire around a suitable nail held in a bench vise. First make up one sample and try it for size at the 10-meter section. To secure the anchor to the spreader, make two turns around the anchor point and then twist the remaining ends into a pigtail. After that step, insert a nail through the loop ID and apply an additional half turn more twist. The critical anchors are those holding the number-14 (1.6mm) horizontal wires. To prevent slippage, number-6 (3.5mm) panhead screws are driven alongside of the wrapped wire; a number-39 drill (0.1 inch or 2.5mm) is used to start them in the bamboo.

Wire stringing is started with the outer number-18 (1.0mm) copperweld vertical sections. Fashion a 9 foot, 2 inch (2.79m) gauge stick from 1x2 inch (25x50mm) wood, and insert it between the inner side of the end anchors before installing the wires and insulators. Leave several inches of pigtail on these wires for connection to the others; do not rely on the anchors for continuity! Transfer the gauge stick to the other side and repeat the procedure.

Next, attach the number-14 (1.6mm) twenty-meter lower wire to the mounted insulator, followed by the upper wire. Pull them up taut, but before securing, sight along the diagonal spreaders and make any corrections necessary. Follow with the number-14 (1.6mm) wires for 10 and 15 meters and the remaining number-18 (1mm) vertical runs, all of which are separate wires to permit proper tension adjustment. The shorting bar and jumper positions can be interchanged for best accessibility. The reflector element shown in

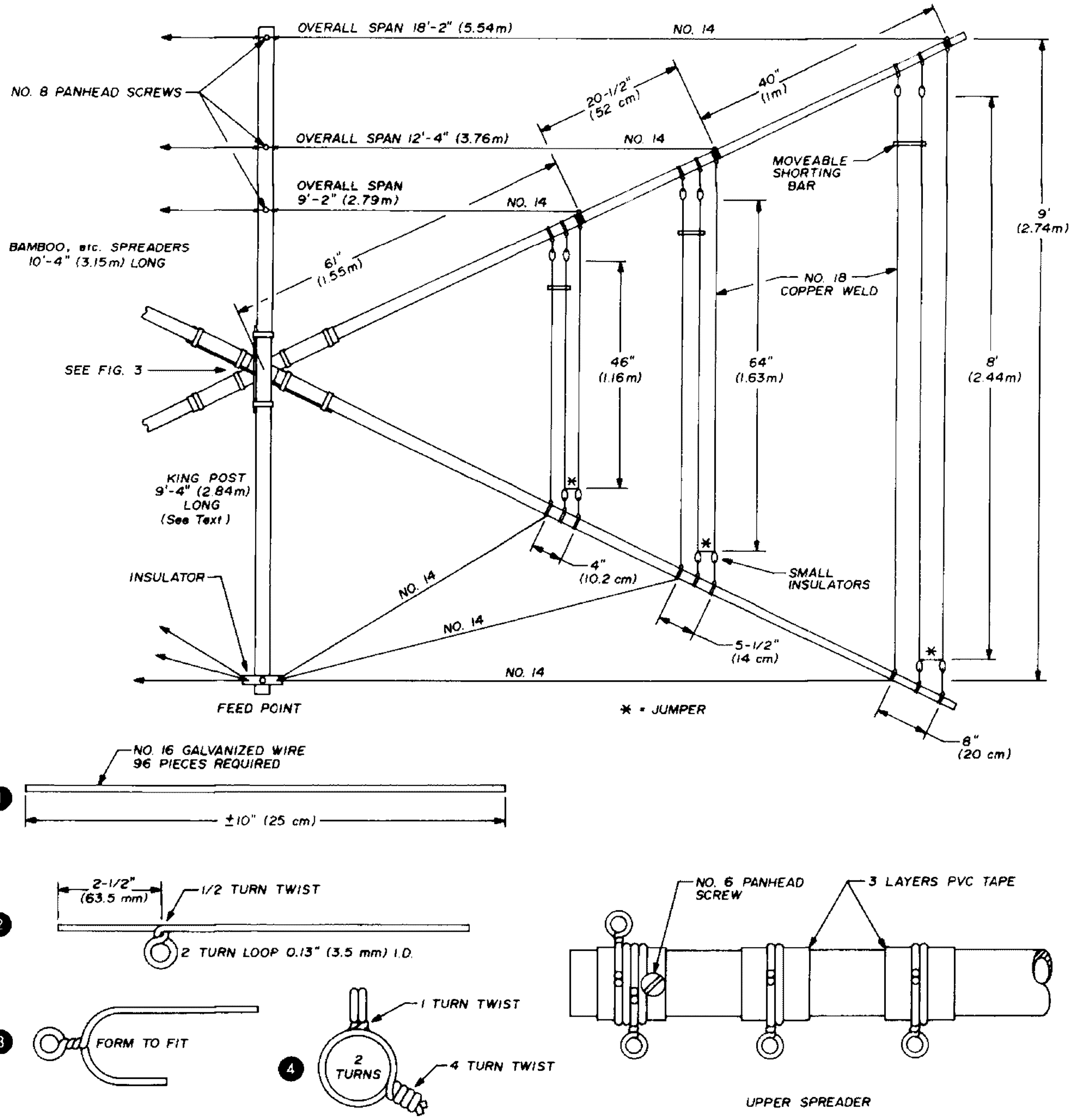


fig. 2. Construction of the reflector and wire loop anchors used in the three-band quad.

fig. 2 is assembled in the same way but without the insulators and shorting bars. To retard corrosion at the anchor loop connectors, dab on a bit of axle grease.

feeding and adjustment

The present direct feed with a single RG-8/U 50-ohm coaxial line is a quick and easy way to get going, but it is probably not the ultimate matching method. A triple gamma match is now feasible since there is now a convenient mast to support the required components. Some improvement was also noted when the original feedline length was extended from 80 feet (24.4m) to 91 feet (27.7m), a length which corresponds closely to multiple of a half wavelength on all three bands. Two other good choices for feedline length are 45 feet (13.7m) and 137 feet (41.8m). Adding the 10- and 15-meter elements to the original 20-meter LPQ lowered its resonant frequency by a

substantial amount, and it required retuning. Tuning is accomplished by adjustment of the shorting bars.

As shown in **fig. 1**, moving the shorting bars upward increases the loop length and lowers the resonant frequency while moving them downward will raise the resonant frequency. Initial

Front-to-back ratio and comparative field strength were checked with a simple detector consisting of a short dipole and a 1N34 diode connected via twisted pair to a 1 mA indicator in the shack. When you are satisfied with the tuning, a fixed jumper should be soldered just below every shorting bar for insurance.

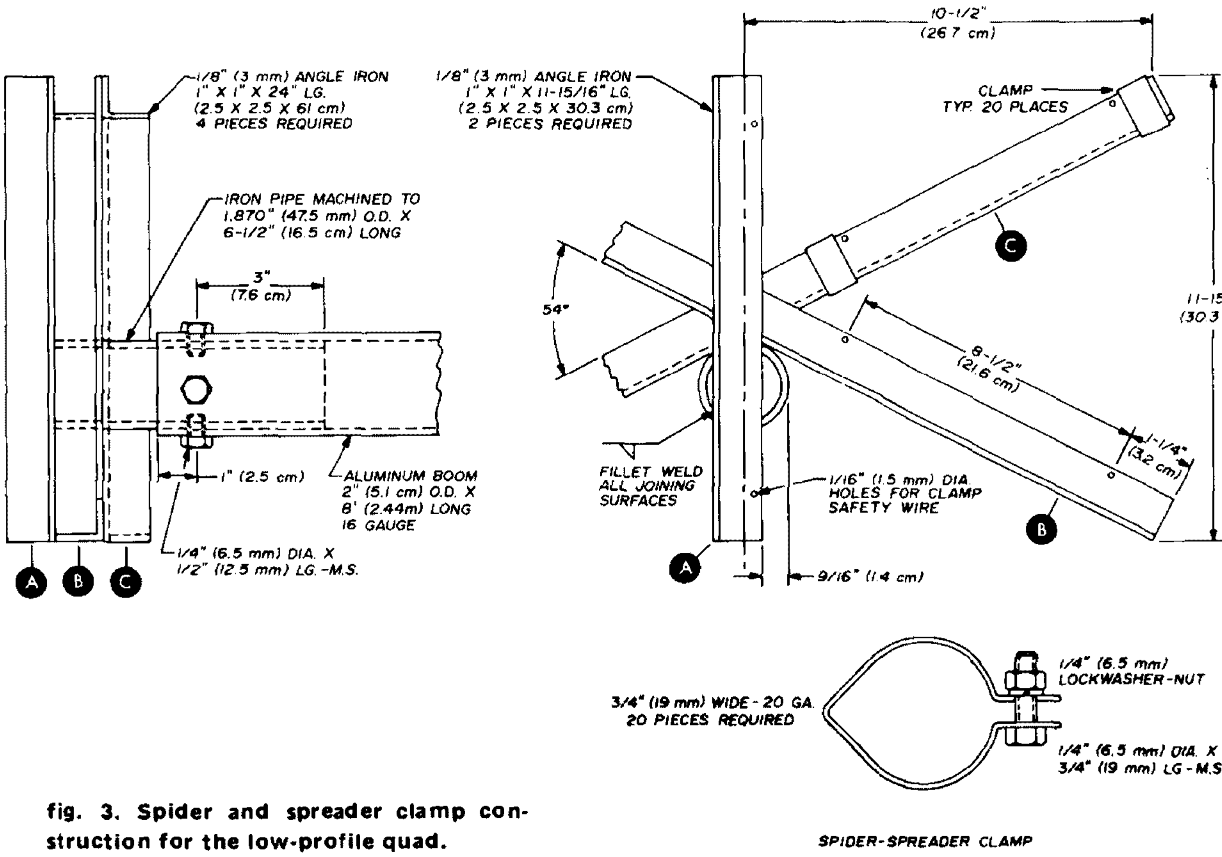


fig. 3. Spider and spreader clamp construction for the low-profile quad.

changes can be made in 5 inch (13cm) increments until the best swr is bracketed. The procedure I used was to raise the antenna, apply low power, and take swr readings across all three bands. At first do not concentrate on trying to obtain the optimum match for any one band. Since three antennas are connected to a common feedline, it may be necessary to make changes to all antennas and repeat the swr checks several times. An antenna impedance bridge will also show frequency changes and may be useful in the tune-up process. The reflector should not require any attention since it is broadly tuned.

After finishing this antenna, another modification to improve the mechanical rigidity of the quad occurred to me: Why not have the 10- and 15-meter loops on the radiator and reflector fastened to the king post at the same point as the 20-meter loops? The elements would then assume a hexagonal shape and the assembly should be even more rigid.

reference

1. J.P. Tyskewicz, W1HXU, "The Low-Profile Quad Antenna," *CQ*, February, 1974, page 24.

ham radio

phase modulation

principles and techniques

When phase modulators are used in vhf fm systems, frequency-shaping circuits must be tailored for maximum intelligibility

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In discussions with other radio amateurs, and in reading such references as the ARRL *Radio Amateur's Handbook* and *FM and Repeaters* manual, I have found that considerable confusion exists concerning modulation standards for vhf-fm. Both of these books explain the difference between fm and phase modulation (pm), but neither point out that it is pm that is normally used on vhf. If homebrew receivers and transmitters are used which do not take this into account, intelligibility will suffer greatly.

fm vs pm

In practice the most obvious difference between fm and pm is in the audio-frequency response. In a pm signal the deviation is proportional to the modulating frequency while in an fm signal it is not. When analyzed mathematically a phase modulator differentiates the audio signal. A pm signal, therefore, has a deviation which rises at 6 dB per octave, so a pm receiver must have a frequency response which falls at 6 dB per octave or the audio will have excessive treble response and be "tinny" sounding. An fm signal received on such a receiver will sound very muffled.

The methods for achieving the desired response are simple and have been reported before,[1] but perhaps a recap is in order. If a phase modulator is used on a multiplier stage of the transmitter (the usual commercial practice in vacuum-tube rigs), the audio response will automatically be correct; hence the only processing necessary is to limit the audio bandwidth and maximum deviation, as described later.

pre-emphasis

The usual practice in homebrew fm equipment is to modulate a crystal oscillator with a varactor diode — a process which is simple but which results in a flat rather than a rising frequency response. Measurements on my Pip-Squeak[2] with an audio generator fed directly into the varactor decoupling network (bypassing the audio stages) show that response is flat from dc to over 20 kHz. Using the original audio circuits, the audio sounds very muffled

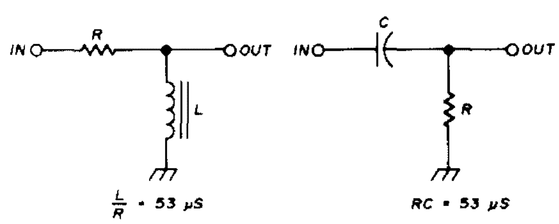


fig. 1. Pre-emphasis networks. For communications circuits with 3 dB rolloff at 3000 Hz the time constant should be 53 microseconds.

unless an inexpensive microphone with peaky high frequency response is used. A simple RC or RL pre-emphasis network, as shown in fig. 1, will cause the audio response to rise at the desired rate.

In this network R and C may have any value provided that their time constant (product) is 53 microseconds. This value is chosen to give 3 dB rolloff at 3000Hz ($R = X_C$ at 3000 Hz). In the RC network shown, R must include the input resistance of the following stage as a parallel component. Similarly, in the RL circuit, R must include the output resistance of the preceding stage as a series component. (The 75-microsecond time constant mentioned in the ARRL books applies to broadcast fm and is used for noise reduction.)

The response curve for the pre-emphasis network and formulas for its

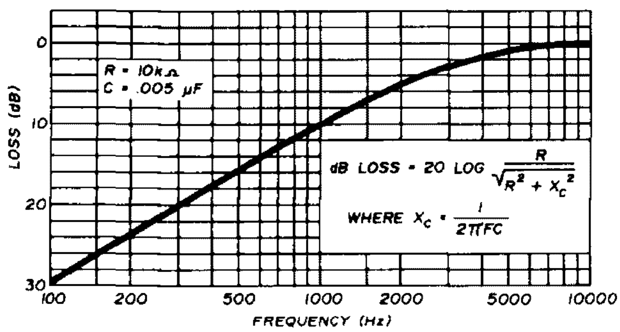


fig. 2. Response of pre-emphasis network. Changing the RC product moves the curve to the left or right but does not change its shape.

calculation are shown in fig. 2. Changing the RC product simply moves the entire curve to the left or to the right but does not change its shape or slope. The curve for the RL network is identical.

clipping and filtering

Good amateur practice (and regulations) require that maximum deviation be limited to ±5 kHz, ±7.5 kHz, or ±15 kHz, depending on the system. Values exceeding these amounts will cause distortion, squelch closing on voice peaks, and unnecessary bandwidth. Since pm deviation increases with both frequency *and* modulation amplitude, it is necessary to restrict both peak amplitude and high-frequency response in pm systems.

A simple shunt clipper is shown in fig. 3. In this circuit resistor R may be simply the output impedance of the driver, but it should be very high compared to the forward resistance of the

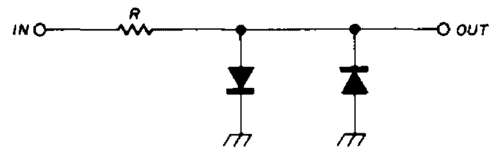
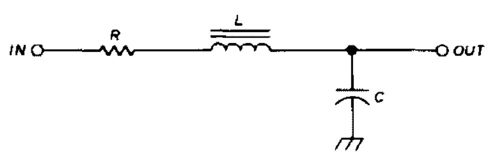


fig. 3. Simple shunt clipper circuit for limiting peak amplitude response of pm systems.

diodes. The diodes may be either germanium or silicon, depending on the audio level. Since the clipping circuit will produce harmonics of the modulating frequency, it should be followed by a lowpass filter.

Level controls should be located on both sides of the clipper. The input control is called a *modulation* control and sets the amount of clipping present on the signal. The output control is a *deviation* control and sets maximum deviation. Insufficient clipping will result in a low audio level; any attempt to correct this by turning up the deviation control will result in excessive peak deviation.

Many commercially built rigs lack a modulation control and depend on carefully matched microphones and audio circuits to set the correct level into the clipper. When designed for mobile use, they often have insufficient gain for normal speech in a quiet fixed station. If your rig suffers from low audio with



$X_L = X_C = R$ AT CUTOFF FREQUENCY

fig. 4. Lowpass LCR filter provides rolloff of 12 dB per octave. Response is plotted in fig. 5.

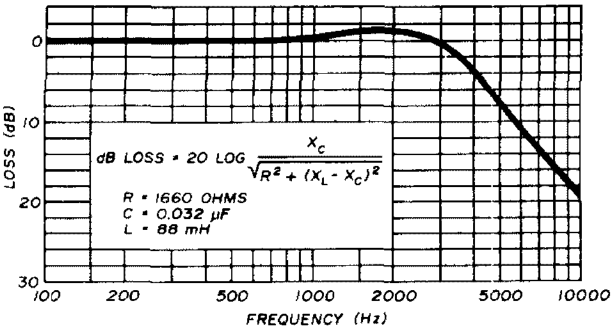


fig. 5. Response of lowpass filter (fig. 4) where R = 1.66k, C = 0.032 µF and L = 88 mH. Rolloff is 12 dB per octave.

the deviation properly set, add a variable gain microphone preamp, don't turn up the deviation.

While a simple RC lowpass filter could be used after the clipper, twice as much rolloff (12 dB per octave) is possible with an LCR filter such as that shown in **fig. 4**. The R, L and C components may be selected by noting that at the cutoff frequency, $X_C = X_L = R$. If L is an 88-mH toroid and the cutoff frequency is 3000 Hz, then C = 0.032 µF and R = 1659 ohms (either 1.5k or 1.8k should be satisfactory).

If R is too high, a peak will develop in the response at the cutoff frequency; if it is too low the high-frequency response will suffer. R includes the output resistance of the preceding stage, and since it must have quite a low value, an emitter follower is suggested.

Since the input impedance of the LCR filter varies widely with frequency, it should not be used to terminate the pre-emphasis network. The stage following the filter must have an input resistance much higher than X_C at the cutoff frequency, say 100k in this case.

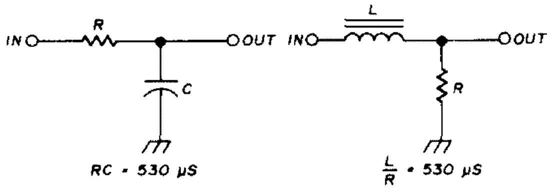


fig. 6. De-emphasis networks are used in phase-modulation receivers to provide 6 dB per octave rolloff. For communications circuits with a low-frequency cutoff of 300 Hz, a 530 microsecond time constant is chosen. Frequency response of the de-emphasis circuit is plotted in fig. 7.

receiving

In the receiver in a pm system the discriminator must be followed by a network giving 6 dB per octave rolloff above 300 Hz. Such a circuit is called a de-emphasis network, and examples are shown in **fig. 6**. If $R = X_C$ (or $R = X_L$) at 300 Hz, a time constant of 530 microseconds results, and the frequency response is that shown in **fig. 7**. Again, changing RC will move the plot to the left or right, but will not change its shape. In mathematical terms, the de-emphasis network integrates the audio signal.

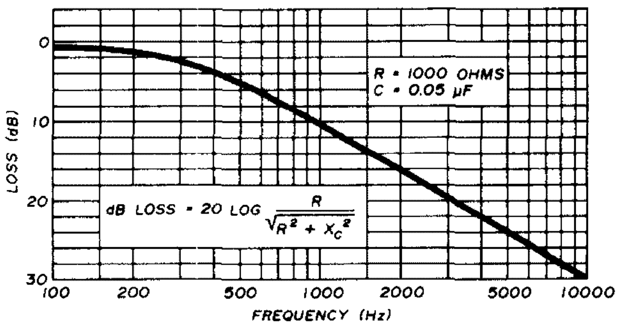


fig. 7. Frequency response of an RC de-emphasis network with R = 10k and C = 0.05 µF.

The block diagrams shown in **figs. 8** and **9** illustrate how the principles discussed may be included in transmitters using either a phase modulator (**fig. 8**) or a frequency modulator (**fig. 9**) to give pm in either case. Clipping is always done after pre-emphasis so that frequencies below 3000 Hz may achieve higher deviation levels than would otherwise be possible.[1]

other techniques

Operational amplifiers may be used as differentiators and integrators, and they have the advantage over passive networks of very low output impedance and a gain of unity or greater rather

than 10 dB loss at 1 kHz. In addition, the frequency response curves are quite linear. A differentiator will serve as a pre-emphasis circuit, and an integrator as a de-emphasis circuit. Suitable circuits are shown in **fig. 10**. Internally compensated op amps such as the μA741 should be used. The circuit of **fig. 10A** may be used in a transmitter and that of **fig. 10B** for a receiver. Parts

resistor. Some receivers, such as the Motorola Sensicon series, have the test point heavily bypassed for both audio and rf. Such bypassing should be reduced sufficiently to allow audio signals to pass. Calibrate the scope by feeding in signals of −10kHz, on frequency, and +10 kHz. Peak deviation may now be read by simply looking at the audio waveform.

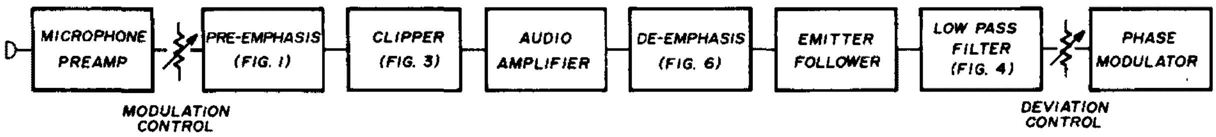


fig. 8. Basic phase modulator system. This is the system most often used in vhf communications equipment.

values are fairly critical to assure correct frequency response. The response curves of these circuits are shown in **fig. 11** (these circuits originally appeared in reference **3**).

measurements

Checking frequency response and distortion on an fm or pm transmitter is easily done if you have access to a few items of test equipment. You will need a good quality wideband fm receiver capable of receiving the transmitted frequency, a dc oscilloscope, and an accu-

In a pm system, only a sine wave is transmitted unchanged; a square wave is sent as a series of positive and negative spikes which would be infinite if not filtered. When clipping begins, therefore, large spikes will appear on the waveform. The peaks of these spikes represent peak deviation and must be set to ±5 kHz, ±7.5 kHz, or ±15 kHz. If they seem out of proportion, the lowpass filter isn't doing its job.

These terrible looking waveforms are restored to their original shape by the de-emphasis network in the receiver. If

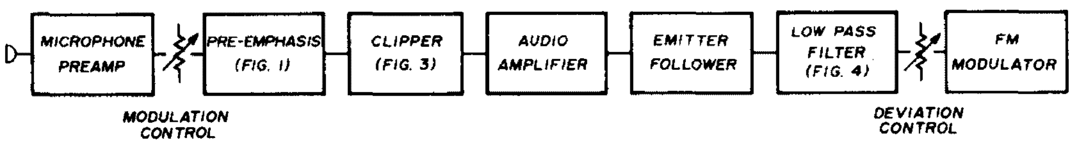


fig. 9. Frequency modulator is similar to pm system but does not require a de-emphasis network.

rate, stable signal generator (a BC-221 is ideal).

First align the discriminator of the receiver very carefully so as to obtain a linear curve of voltage vs frequency. Connect the oscilloscope to the discriminator test point. If no test point is provided, connect the scope directly to the discriminator through a 1 megohm

you are not convinced, either put a de-emphasis circuit on the input of the scope, or connect the scope to the speaker terminals instead of the discriminator.

Using this method will show up one fault of a varactor diode modulator: the modulation will be asymmetrical, going more in one direction than the other.

This is partly because the capacitance-vs-voltage curve of such a diode is not a straight line. Fortunately, most voice modulation is below the level at which the distortion becomes obvious, so it usually sounds all right.

repeaters

The frequency response of a repeater should be as flat as possible over the desired bandwidth of 300 to 3000 Hz. In fact, it may be desirable to extend the low-frequency response to 100 Hz or so if the use of sub-audible tones is contemplated. This may be accomplished by increasing the time constant, RC, in **fig. 6** to 1.6 milliseconds. No attempt should be made to tailor the response for better "communication quality" at the repeater, since the users' rigs will already include such provisions and doing it twice will make things worse. Ideally, the repeater output should be identical to its input in all respects within the limits of audio bandwidth and deviation.

The frequency response of some commercial rigs is less than ideal, and before using them as a repeater they could benefit from the application of some of the points mentioned in this article. When the repeater is installed, a careful set of measurements should be made to ensure that the following conditions are satisfied:

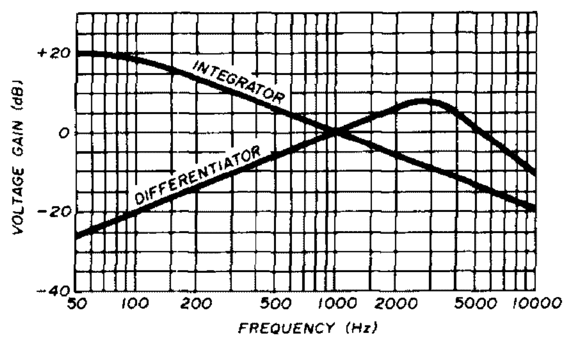


fig. 11. Frequency response of op-amp integrator and differentiator circuits shown in fig. 10.

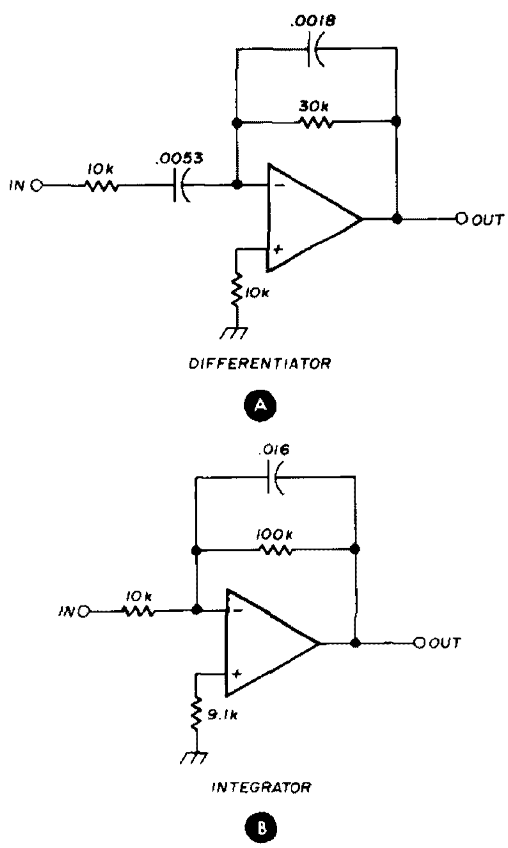


fig. 10. Operational amplifiers may be used as differentiators (pre-emphasis) or integrators (de-emphasis) as shown here. Frequency response of these circuits is shown in fig. 11.

1. The input and output deviations must be identical up to the maximum chosen by the system.

2. Inputs with greater deviations than this maximum should be clipped to the maximum.

3. The frequency response should be flat within ± **2** dB from 300 to 3000 Hz for signals below the clipping level.

4. There should be little distortion on *signals below the clipping* level.

Here is the method I use to make these tests: The equipment is set up as shown in **fig. 12**. The receiver must be switchable between the repeater input and output frequencies, and separate receive and transmit antennas must be used (unless you have the facilities to work duplex on one antenna). If low power is used, and the repeater signal is

fairly strong, full duplex operation should be possible. Turn your transmitter deviation all the way up to essentially remove the clipper from the circuit. Calibrate the scope as described earlier. Set the repeater deviation and modulation controls at maximum. The following steps are shown for a ±5 kHz system, but may be scaled for any desired deviation.

5. Feed a **1**-kHz tone to the transmitter and adjust the audio level for ±7.5 kHz deviation using the receiver and scope on your own frequency. Switch the receiver to the repeater frequency and have the repeater deviation reduced to ±5 kHz.

6. Feed a 1-kHz tone into the transmitter and set the deviation to a low value, say ±2.5 kHz. Have the repeater modulation control adjusted for an output deviation of ±2.5 kHz. Make sure that the audio waveshape on the scope is an undistorted sine wave.

7. Repeat the above checks at different levels and frequencies to make sure that conditions **1** and **2** above are met.

8. Feed in tones from 300 to 3000 Hz, setting the transmitter deviation to ±2.5 kHz each time, and see that the output deviation does not vary by more than ±2 dB (use an audio vtvm in parallel with the scope if desired).

Note that the repeater receiver and transmitter may both be fm instead of pm if desired, as the system frequency response will be flat in either case. **Fig. 13** shows the overall frequency response using the pre-emphasis and de-emphasis circuits mentioned earlier. If a flatter response curve is required, consideration should be given to using the op-amp integrator and differentiator circuits described previously.

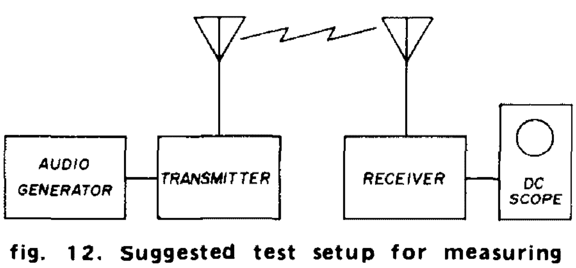


fig. 12. Suggested test setup for measuring frequency response of a repeater (see text).

conclusions

The fm systems used on vhf are actually pm, not fm. Many homebrew transmitters produce fm, not pm, but the required conversion can be effected by simple RC networks. Modulation and deviation are not the same thing.

If care and attention are paid to audio circuit design for vhf fm transmitters and receivers, the resulting voice quality is almost good enough to be called hi-fi while still retaining full communications effectiveness, and a well designed repeater can re-transmit a signal with no noticeable change in this good quality. First-class audio is not hard to obtain — let's hear more of it.

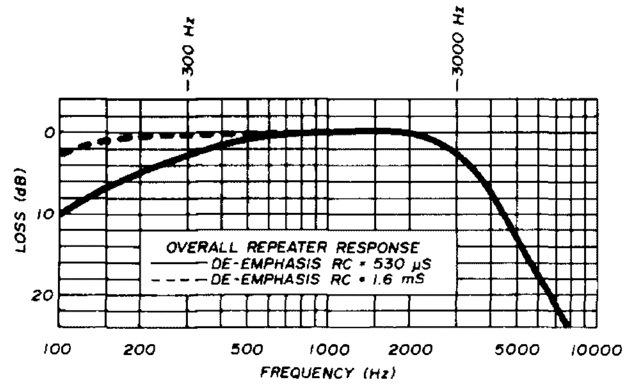


fig. 13. Overall frequency response of a repeater using the pre-emphasis and de-emphasis circuits shown in figs. 1 and 6.

references

1. Les Cobb, W6TEE, "Modulation Standards for VHF FM," *ham radio*, June, 1970, page 16.
2. Doug DeMaw, W1CER, "An FM Pip-Squeak for 2 Meters," *QST*, March, 1971, page 21.
3. "An Applications Guide for Operational Amplifiers," Application Note AN-20, National Semiconductor Corp., Santa Clara, California, February, 1969.

ham radio

high performance sync generator for amateur television

High performance television sync generator uses National MM5320 IC to provide all pulse signals needed for both black-and-white and color TV

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Most amateur television installations tend at first to be derived from low-cost surveillance TV cameras. These cameras use a random interlace system and often combine both blanking and sync as a single pulse. Professional TV systems, however, use a very stable sync generator which will interlace the scan pattern for maximum definition. They also provide a separate blanking pulse group which will give a clean picture with sharp edges and no retrace problems. The final addition to a professional sync generator is color synchronizing capability. Now you can get all this by using one special purpose integrated circuit plus some peripheral ICs and other parts.

circuit

The new National Semiconductor MM5320 mosfet circuit is the heart of this sync generator, and it will do everything required of an amateur TV sync generator and more. You start by generating a crystal controlled 14318-kHz signal and divide it by seven to give 2045 kHz. This is the basic clock frequency which drives the MM5320 and in turn produces all the correct sync pulses, both horizontal and vertical, equalization and serration pulses, blanking, drive signals and color gate plus a few other things you may be interested in if you own a video tape recorder and other exotics.

I expect that some amateurs will want to try color television now that the one-tube color camera has been developed. Thus I have included the color parts of the sync generator in the schematic. If you don't need them, just leave them out. Basically the color subcarrier frequency of 3579 kHz is generated by dividing the crystal frequency down 4 times. If you do use color, you may want to add a trimmer capacitor to the crystal oscillator so you can adjust this signal to the exact frequency. I have

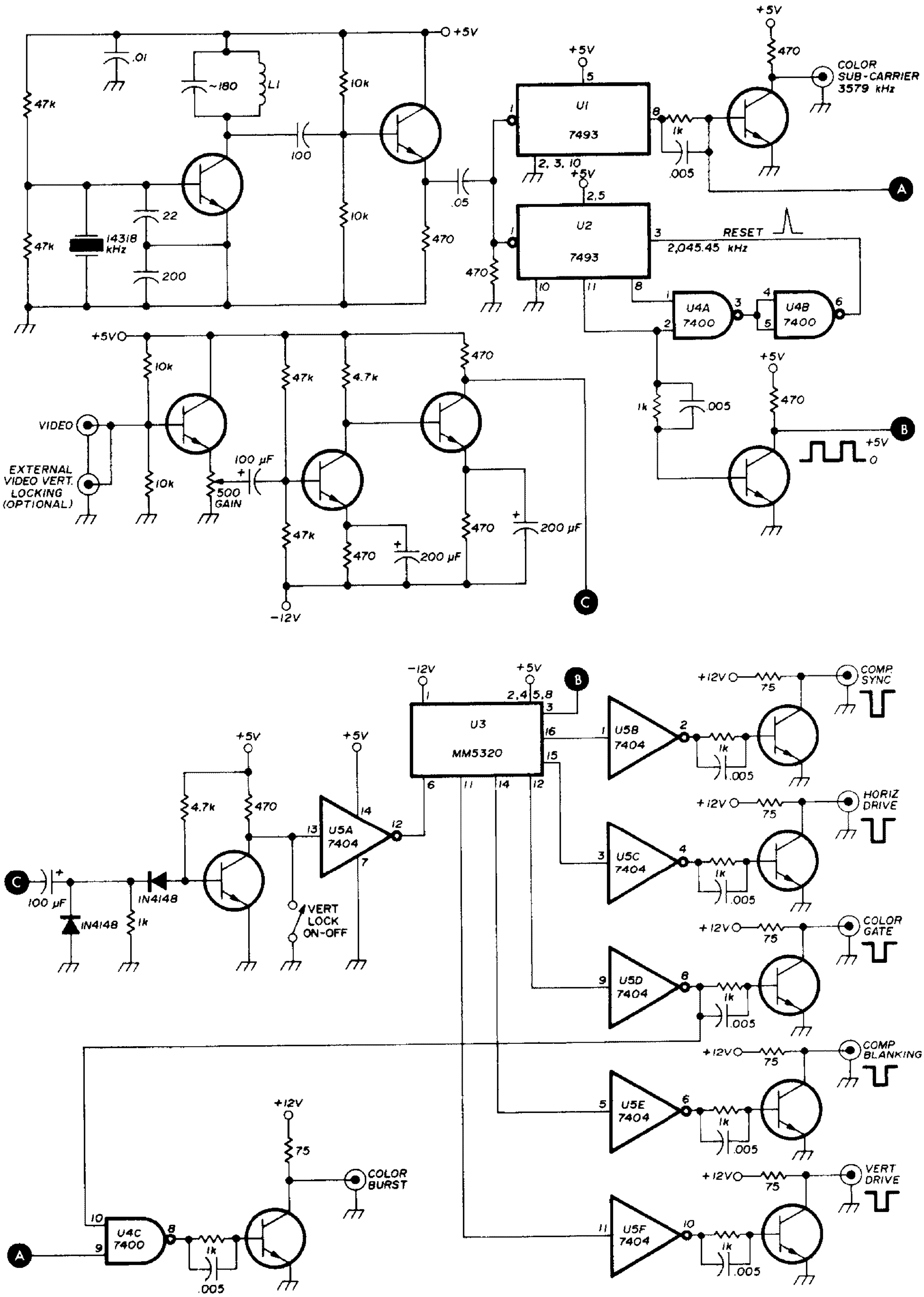

fig. 1. Schematic diagram of the television sync generator. A 16-pin socket is recommended for the National MM5320 IC (see caution in text). L1 is 25 turns no. 24, closewound on a 3/16" (5mm) diameter form.

also included a color-burst output which can be added to your signal just for fun even if you are using black and white (please use a switch so you can remove the signal and keep peace with the guys with color monitors). A color gate signal is also there if you need it.

Notice that all circuits connected to the IC are buffered to protect the chip. In addition, the outputs are driven hard with discrete transistors to provide solid signals for a 72-ohm line.* All signals are negative going (for true) at the outputs.

construction

The transistors I used were all epoxy types, high-speed switches with a voltage rating of 30 volts or better. About five different low-cost types were tried and all worked okay. I did have a MM5320 fail, so I recommend that you use a 16-pin socket for the chip and build everything else on perfboard with number-28 tinned wire and Teflon spaghetti before plugging in the chip. When all the wiring is completed and checked, ground yourself to the chassis and the IC holder to your fingers, then remove the IC from the holder and plug it in. Remember to keep yourself grounded to the chassis with one hand. Don't stroke the top of the IC or you may build up enough static charge to turn the circuit off. Also, if you put a heat dissipator on the IC, ground it too as any charge on it may impede circuit action.

The schematic also shows some vertical sync-locking circuits which may be connected to your video tape recorder to keep the picture from rolling as you switch from camera to recorder. If you don't need this, just connect pin 6 of the MM5320 to +5 volts and forget that part of the circuit leading up to pin 6.

*If coaxial cable is used to carry the signal from the sync generator to other equipment, add resistors in series with each output point shown in the schematic. Use 68-ohm resistors for 70-ohm coax and 47-ohm resistors for 52-ohm coax.

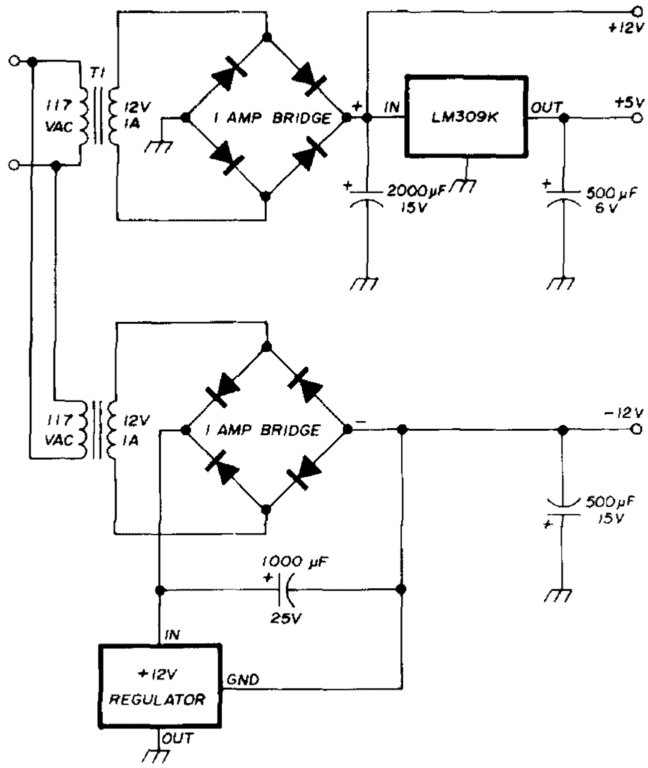

fig. 2. Power supply for the television sync generator uses voltage-regulator ICs.

Now that you have a sync generator right up to network standards (almost anyway), you are ready to convert your old camera over to full 2:1 interlace scan with correct blanking and a stable sync frequency. You will have to figure out the actual interconnections but let me suggest using some high-voltage HEP (or similar) transistors to drive your vacuum-tube sync circuits and get rid of those old free-running oscillators.

As for testing the unit, a frequency check is nice, but is only required for color. Connecting each output to my old Tektronix 310 scope provided nice sharp +5 volt signals which looked just like those out of the $2000 sync generator down at the local TV station. Quarter- or half-watt resistors work well everywhere and only the MM5320 got warm. This was solved by adding about two square inches of aluminum (in the form of an L) to the top of the IC with some five-minute epoxy (and grounding the heatsink).

ham radio

multiplexed seven-segment readouts

Surplus calculator digital readouts may be easily adapted for other uses

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The tremendous volume of low-cost, hand-held digital calculators has led to the appearance of a number of solid-state digital readouts in the bargain lists. Many of these units are internally wired with common cathodes for each digit and all like segment anodes tied together. The units are quite small, but this doesn't seem to be a problem in calculators so they should be usable in frequency counters as well.

The internal connections necessitate multiplexing the readout system. This means that each digit is activated in turn and is only on a small part of the time. In the case of a nine-digit readout, each numeral will be on only one-ninth of the time. If this action is repeated 30 or more times a second, the entire row of nine digits will appear to be on at once.

multiplex circuit

To accomplish the multiplexing, each individual digit's cathode lead is grounded in turn and, at the same time, the common seven-segment anodes are tied to the proper decade through a decoder-driver. The circuit shown in **fig. 1** will do this using easily obtained parts.

Briefly, the circuit works as follows; a 7441 BCD-to-decimal decoder is driven by a 7490 counter operating with a 1kHz input from the frequency stan-

dard countdown chain of the counter. This causes the ten outputs of the 7441 to go low in sequence. Each open collector is "pulled up" by a resistor to the +5 volt supply. As each output goes low, it outputs are connected through diodes to the ABCD inputs of the single 7448 decoder which drives the segment anodes of all the digits thru current limiting resistors.

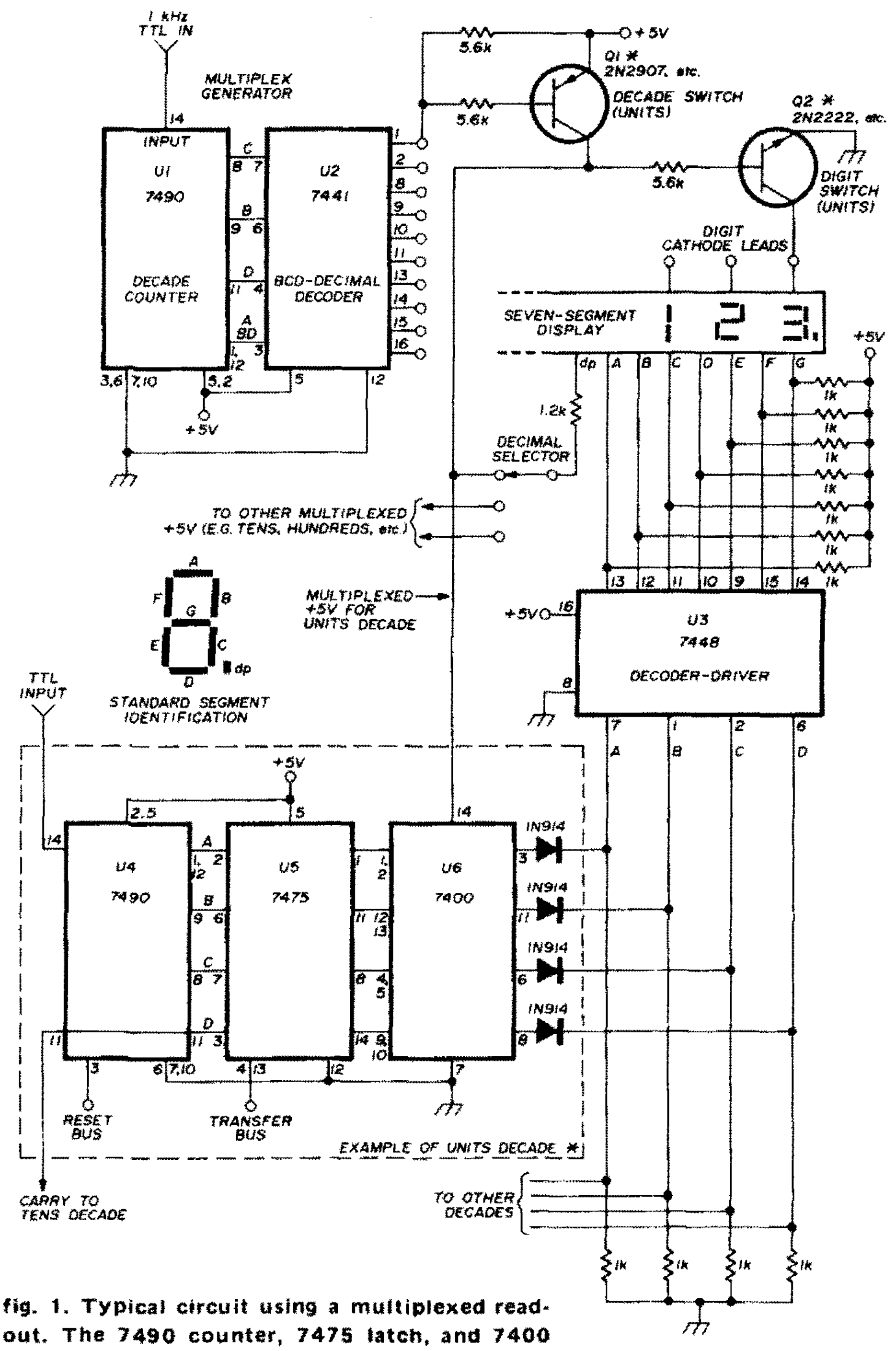


fig. 1. Typical circuit using a multiplexed readout. The 7490 counter, 7475 latch, and 7400 driver ICs and circuitry marked with an asterisk are repeated for each decade.

causes a pnp transistor to switch the five-volt supply voltage of a 7400 tied to the ABCD compliments ($\overline{Q}$) of the 7475 latch in one decade. The 7400's

The 7448's inputs are grounded through resistors which hold these inputs low in the absence of high outputs from the 7400. In other words, the

7400 can only send high signals to the 7448. This allows all the decades to be tied to the same decoder-driver without problems occuring when other 7400 outputs are low since only one 7400 has supply voltage at any given time and is capable of producing a high. The switched supply voltage is also applied to an npn transistor which grounds the common cathode lead of the proper digit.

The 7441 multiplex generator can supply up to ten digits. In the unit I built no attention was paid to the sequence of activated digits, and this was left to be determined for convenience of circuit-board layout.

The readouts I used were advertised as similar to the DL-33 and have a 5 mA per segment rating. Surprisingly, these units have been operating at about 1.6 mA per segment with adequate brightness except when under direct illumination by the light over my workbench. Current measurement was made with the multiplexing off and without the 1000-ohm resistors. Needless to say, this results in very low current drain for the nine digits. Segment current rose to 4.5 mA with the resistors in the circuit, and the brightness increased considerably. With multiplexing, readout brightness and dissipation are reduced because of the low duty cycle, and this should insure long life.

The rest of the circuitry for frequency counters or other applications may follow any of the usual techniques without the display method necessitating any changes.

summary

Although the multiplex system is perhaps more complex, it actually has less circuit repetition. As the number of digits increases, this approach becomes more attractive. Its use should encourage construction of frequency counters with sufficient readout to directly display high-frequency signals with 1-Hz resolution.

ham radio

low-noise uhf preamplifier and converter

A simple, easy-to-build converter and low-noise preamplifier for the 450-MHz amateur band

Jerry Vogt, WA2GCF, Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612

The latest **in** the **Hamtronics** series of receiver-related projects was quite a challenge. I set out to develop a low-noise receiving preamp and companion converter for the 432 and 450 MHz amateur bands and adjacent commercial bands. This article describes the results of several months work. As anyone who has attempted to build solid-state uhf gear knows, layout and construction technique is vital – not only to come up with a good basic design, but to make one which even works!

My goal was to design units which could be duplicated easily, with no special metal-working facilities. I was determined it could be done with a single-sided, printed-circuit board chassis and simple shielding components. In the several months it took, I researched dozens of old magazine articles, handbooks and manufacturer's application notes. I designed and breadboarded at least four different converters and about a dozen preamps. At one point, I even soldered the metal can of an fet directly to the metalization on a feedthrough capacitor to try to reduce lead inductance. That gives you an idea how critical lead lengths are at uhf.

Having tried many semiconductor devices, the one I finally settled on is the new *Siliconix* J308 super-fet. This is an n-channel device similar to two E300 fets in a monolithic parallel combination. It has a very high transconductance (about 13000 μmhos at 450 MHz) and a low noise figure (about 3.4 dB at 450 MHz if correctly matched).

uhf converter

The converter shown in **fig. 1** uses a low-noise jfet mixer, Q1, with a high-Q, two-pole filter input which may be tuned to any frequency in the 400 to 500 MHz range. It is designed to provide optimum performance at low cost for use with receivers operating as an i-f in the range from 30 to 170 MHz. Possible uses include 432-MHz CW, TV, ssb, 450-MHz fm and 450 to 470 MHz commercial fm. The passband of the front end is wide enough to allow operation on amateur TV.

Channel selection, for channelized operation, may be accomplished either at the i-f or by using a multichannel adapter in place of the built-in local oscillator. The converter may be used for scanner operation by switching oscillator frequencies or by using a scanning receiver at the i-f.

Because the converter has a low inherent noise figure, for best results it is recommended that it be used with a sensitive, low-noise receiver. Generally, due to the low-noise design of the converter, the sensitivity will be determined by the sensitivity of the i-f receiver. Note that the converter is not designed to provide additional gain to the i-f receiver. It is accepted practice, in mod-

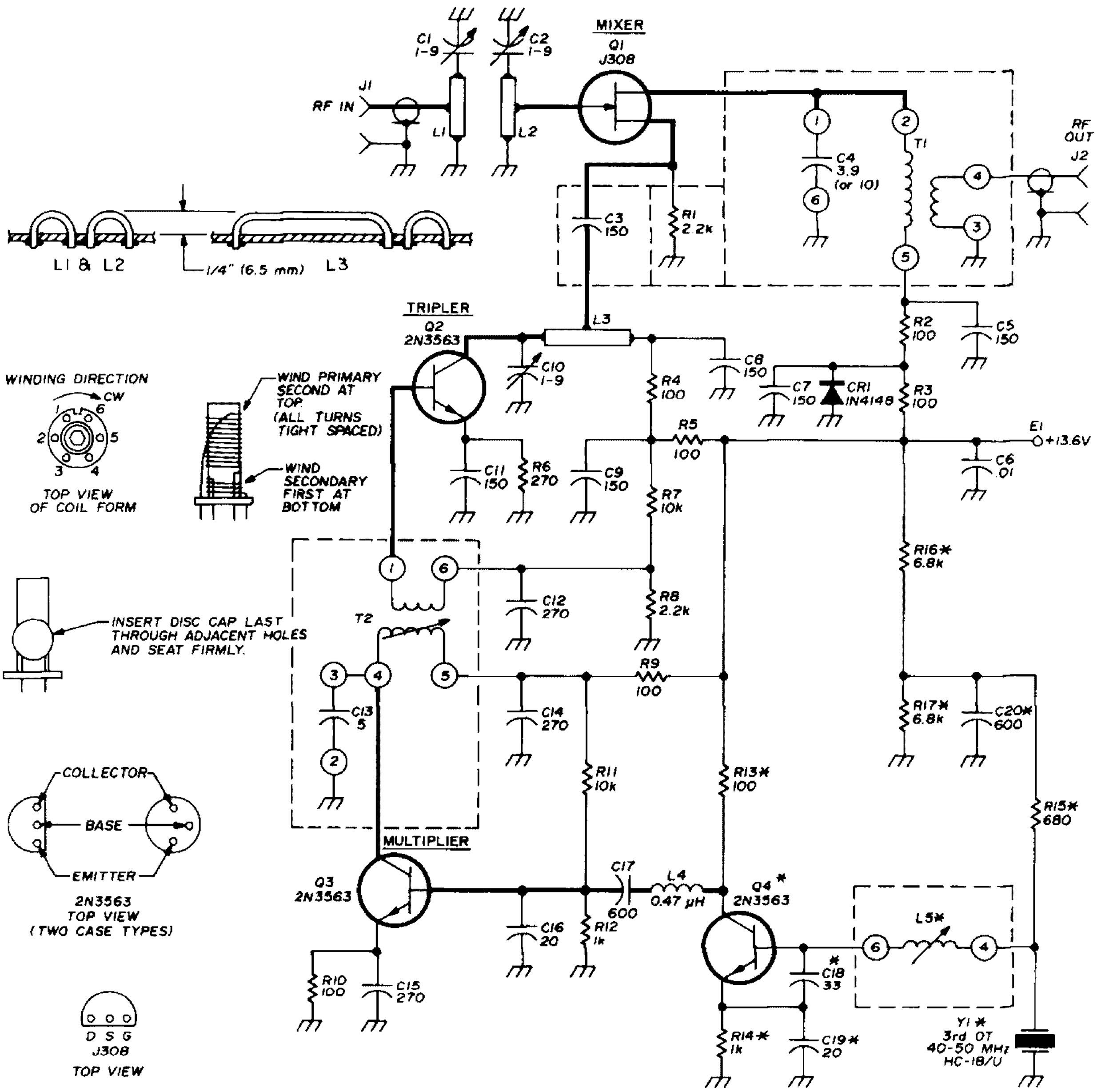


fig. 1. Schematic diagram of the uhf converter. Components marked with an asterisk are not required if the multi-channel oscillator (fig. 7) is used. Inductor L5 is 11-2/3 turns no. 22 (0.6mm) on ¼" (6.5mm) diameter, slug-tuned coil form (carbonyl-J slug). Windings of other inductors depend upon operating frequency and are described in the text. Printed-circuit layout is shown in fig. 2.

ern uhf design, to mix down as soon as possible and use only enough converter gain to offset losses in the tuned filter. This minimizes front-end overload in urban areas with high level, adjacent-band signals. A pre-amplifier may be used between the converter and the i-f receiver if the receiver has insufficient gain. If desired, a uhf preamp such as the one described later in this article can be used.*

*The following kits are being made available in conjunction with this article: U20-450 uhf converter kit, $20; P15-450 uhf preamplifier kit, $15; P25-450 uhf preamplifier, wired and tested, $30; A13-45 uhf six-channel oscillator adapter, $12.95; AS10 scanner adapter with LEDs, $10. Kits are complete except for crystals and include predrilled G-10 printed-circuit boards. Crystal certificates are available for $5.50 each. For more information, send a self-addressed, stamped envelope to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

crystals

The crystals in the converter local oscillator (or in the accessory six-channel adapter) are series-resonant, third-overtone, 0.002% HC-18/U types. The required crystal frequency is equal to the channel frequency minus the i-f frequency divided by six. One kHz is subtracted from the calculated crystal frequency to center the crystal in the trimming range of the oscillator circuit. When the local-oscillator frequency is in the 400-MHz range instead of the 300-MHz range, as used for a high-band i-f, the division factor is nine instead of six. This is necessary to place the crystal frequency in the 40 to 55 MHz range.

When selecting the intermediate frequency, be sure to consider the effects of vhf feedthrough. I did some testing with 146.94 MHz as an i-f, and a few local amateurs had signals strong enough that they could be heard with the vhf antenna disconnected. If you want to use a receiver on an existing channel, choose one of the less popular frequencies.

converter construction

All the vhf coils are wound clockwise, using the solderable wire supplied with the kit (other coil forms and winding techniques may be substituted if you elect to build with your own components. All turns are close spaced. The detail in fig. 1 is exaggerated for clarity, but all leads should be pulled tight. No fancy bends are required. Holes in the base of the coil form are numbered as shown. Thus, coils can be prewound and then installed on the board with the keyway oriented toward the mixer end of the board. Secondaries should be wound first, followed by primaries, followed by insertion of the disc capacitor, if any.

The leads should be inserted through the circuit board while the coil form is spaced slightly away from the board; then the coil form is seated into place.

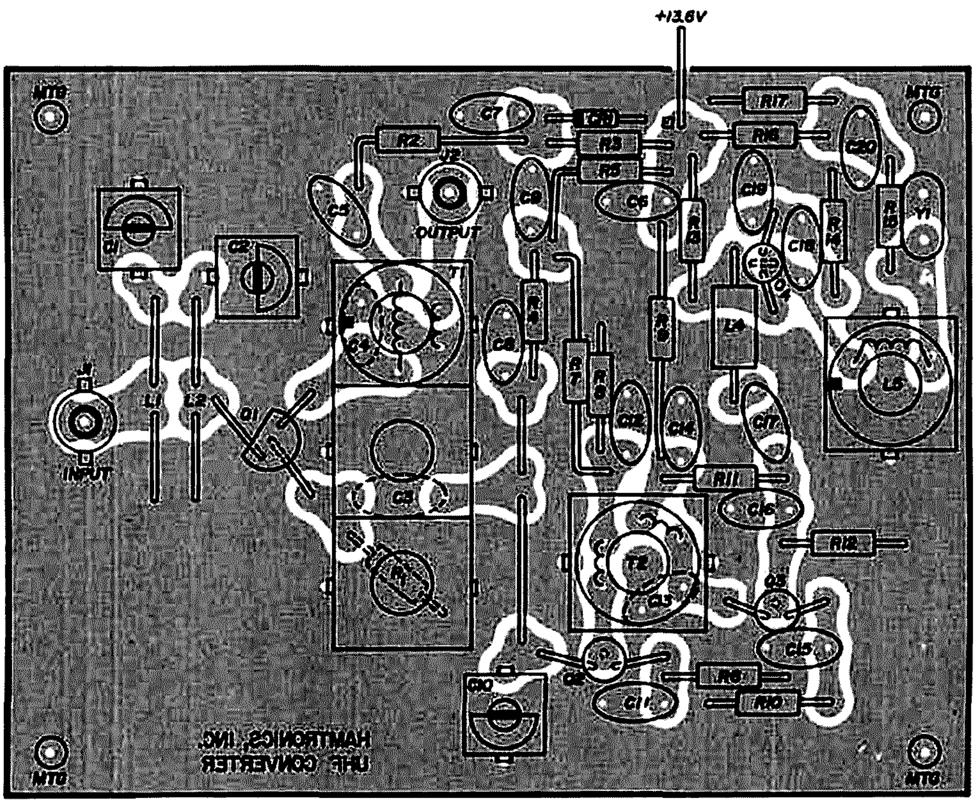
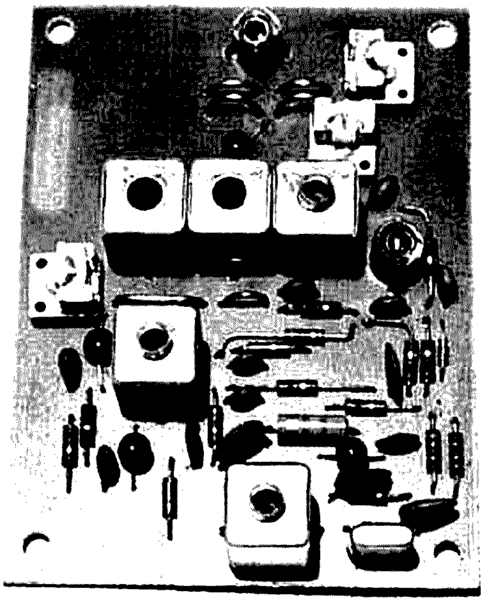


fig. 2. Component layout for the uhf converter. Circuit board is shown in fig. 3.



Basic uhf converter covers the range from 420 to 470 MHz. Printed-circuit layout is shown in fig. 5.

Do not attempt to insert the leads with the coil form flush against board. After all the coils are installed, application of heat from a very hot soldering iron for 10 to 15 seconds (with solder applied) will automatically strip the *Solderon* wire and provide a good solder bond. If you prefer, all leads may be stripped before the coil form is installed on the board. Do not solder-strip the leads unless the coil form is mounted on the board as the leads may migrate into the plastic form.

Coil L5 is always 11-2/3 turns number 22 (0.6mm), with the winding starting at terminal 4 and ending at terminal 6. Construction of transformer T2 depends upon the oscillator injection frequency. For injection in the 300-MHz range the first multiplier doubles to the 90 to 110 MHz range and the secondary of T2 consists of 1-5/6 turn number 22 (0.6mm) from terminal 6 to terminal 1; the primary consists of 4-1/6 turns number 22 (0.6mm) from terminal 5 to terminal 4. A 5-pF capacitor is installed in holes 2 and 3. For mixer injection in the 400-MHz range (low i-f), the primary of T2 consists of 3-1/6 turns number 22 (0.6mm) to allow the multiplier to triple to the 120 to 150 MHz range. The secondary is the same as

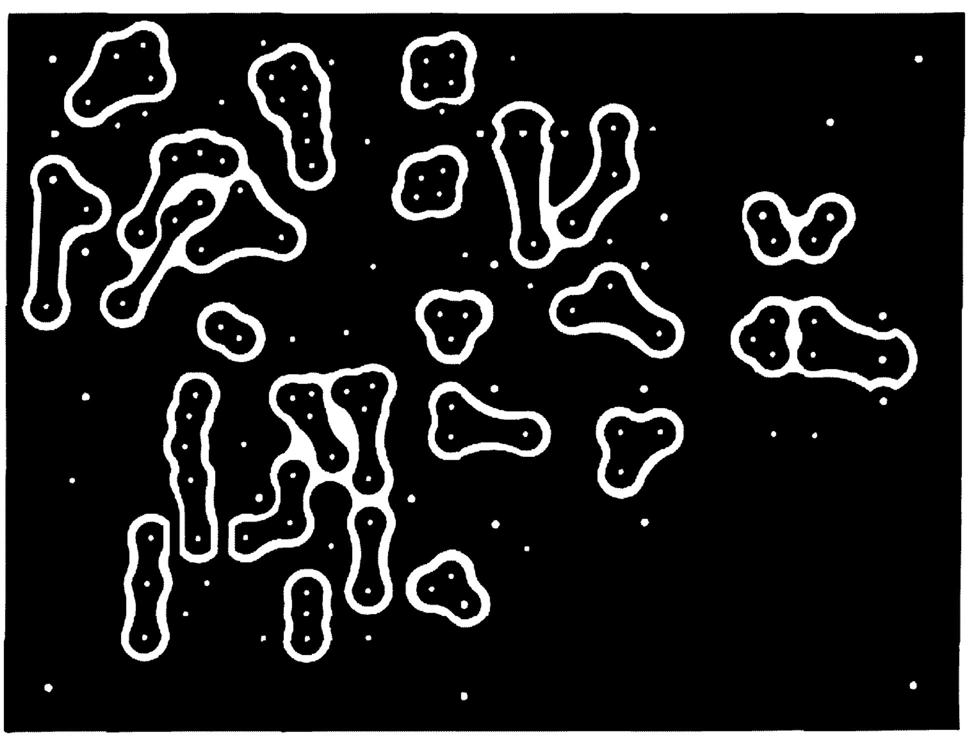


fig. 3. Full-sized printed-circuit board for the uhf converter. Component layout is shown in fig. 2.

above. Other capacitors or numbers of turns may be used to provide resonance for other local-oscillator injection frequencies.

The secondary of transformer T1 is wound from terminal 4 to terminal 3. The primary is then wound from terminal 2 to terminal 5 (all windings number 22 [0.6mm] wire). The capacitor is installed between 6 and 1. For an i-f in the range of 140 to 170 MHz, the

connection to the board. The variable capacitors will tune over a wide range to compensate for coil variations. Excess leads should be trimmed on the foil side of the board after soldering the number-18 (1mm) wires to the board.

connectors

The popular RCA style phono connectors are used for coaxial cable connections to the circuit board. However,

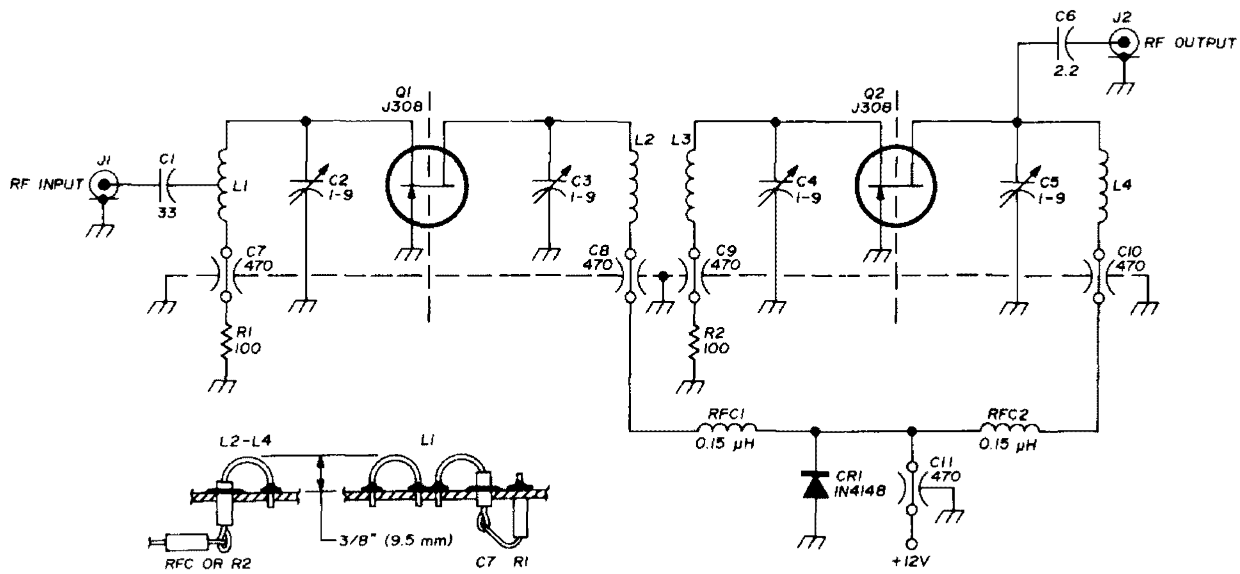


fig. 4. Two-stage uhf preamplifier provides 15 to 25 dB gain. Capacitors C2-C5 are vertically-mounted E. F. Johnson 9-pF air variables. Printed-circuit layout is shown in fig. 5.

secondary should be 1-1/6 turn; primary, 5½ turns; and capacitor, 3.9 pF. For an i-f in the range of 40 to 60 MHz, the secondary is 3-1/6 turns; primary, 15½ turns; and capacitor, 10 pF. For an i-f of 20 to 30 MHz, secondary is 4-1/6 turns; primary, 21½ turns; and capacitor, 10 pF. Note that further adjustment of the number of primary turns may be necessary for other frequencies. Smaller wire sizes may be desired at large numbers of primary turns so that this winding is near the lower half of the coil form.

Coils L1, L2 and L3 are formed as shown in **fig.** 1, using number-18 (1mm) tinned bus wire. Forming is not critical because the coils are essentially straight pieces of wire which are bent to allow

any connectors used in the line at uhf should be a constant-impedance type (such as type-N) for low loss; phono connectors and uhf-type coaxial connectors put a "bump" in the line in such applications. Likewise, the coaxial cable type should be carefully chosen for low loss at uhf. RG-58/U coax, for example, has too high loss for long runs (it may be okay for short runs in mobile operation or for monitor applications with strong signals). If a transmitter is involved, a good coaxial relay should be used, both to minimize signal loss and to prevent coupling of large amounts of rf into the front end of the converter.

converter alignment

The most difficult part of the align-

ment process is obtaining a stable test signal. Even my HP-608 signal generator takes several hours to settle down enough to stay within a 5-kHz passband at uhf. The best solution is to use a weak-signal source such as one of those described in the amateur magazines.[1,2] An on-the-air test is an alternative if you can find an appropriate signal.

Start with all adjustments at about mid-range. Tune in the test signal and peak all adjustments. If T1 or T2 do not peak within the range of their tuning slugs, an adjustment in the number of primary turns may be necessary. Then adjust the oscillator trimmer coil, L5, to net the converter to the channel frequency by monitoring the receiver discriminator or S-meter. Note that the crystal may be pulled enough to adjust over a range of about 10 kHz at uhf. The extra 1 kHz which is subtracted during the crystal calculation centers the oscillator adjustment range.

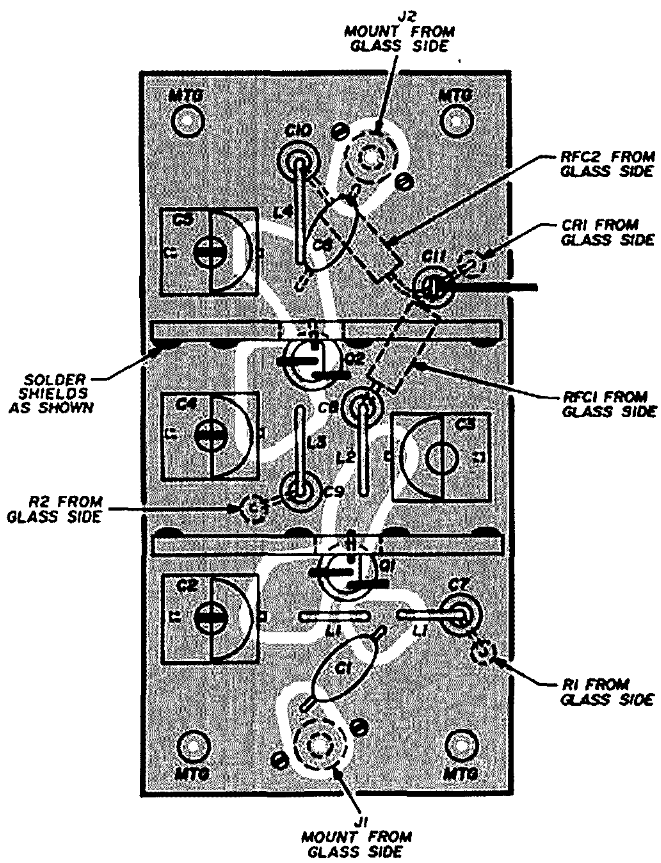


fig. 5. Component layout for the uhf preamplifier. Circuit board is shown in fig. 6.

The dc voltages shown on the schematic diagram are a guide, and are based on using a 13.5 volt power supply. Probably the most common trouble, based on my experience with vhf preamps, is a burned-out fet caused by excessive rf from a transmitter or transients on the B+ line from relay coils, etc. Diode CR1 provides protection from reverse transients, but forward voltage spikes from faulty power supplies may still get through. Remember, if you encounter problems during initial tests, it is very easy to install components in the wrong places or to make a cold solder joint. Double check every connection!

uhf preamplifier

The uhf preamp shown in fig. 4 is a two-stage grounded-gate device which provides a gain of 15 to 25 dB. Although designed primarily for use on the 432 and 450 MHz amateur bands, the preamp may also be built for use on the 300-MHz aircraft band, the 470-MHz commercial band, or the lower vhf TV channels. Its passband is sufficiently wide for amateur TV or selected commercial TV channels.

The preamp components are mounted on both sides of the circuit board to provide the high degree of shielding required at uhf. Critical components are installed on the copper side of the board with absolutely minimum lead lengths. Feedthrough capacitors interconnect components on the other side of the board to provide thorough bypassing. When construction is almost finished, shield plates are installed between the input and output of each stage to prevent feedback. Following is a step-by-step summary of construction:

1. Install the coaxial connectors from the fiberglass side of board by gently rocking them into place. Spot solder the four tabs on the copper side.

2. Spot solder the feedthrough capacitors and variable capacitors in place as shown in the pictorial, fig. 5.

3. Cut number-18 (1mm) wires, form as shown in fig. 4, and spot solder them to the feedthrough capacitors and the circuit board. Loops on feedthrough can be cut off and solder can be applied directly to metalization on the ceramic. For each coil, first apply solder to the feedthrough or one pad of the board. Then solder that side of coil in place while holding the coil loop with pliers. Straighten as required and then solder the other end of the loop. Coils L2 and L3 should lean toward each other for tight coupling, about 1/8-inch (3mm) apart at top of loop (not shown).

4. Bend the leads of the disc capacitors away from the body of the capacitor and trim as shown. Solder short remaining leads to strip pads on the circuit board.

5. Spot solder the shields (made of 1½x1 inch [38x25mm] G-10 board material) to ground areas, centering a small notch over drain lead pad area. This notch must be very small to prevent feedback. Realign the shield after tacking it in one place with solder; then solder the shield firmly in place.

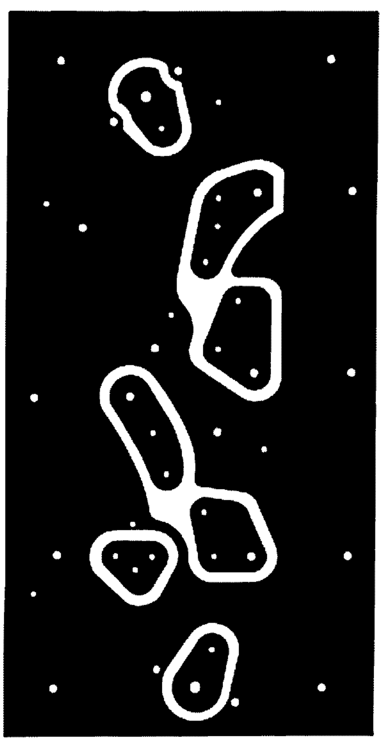
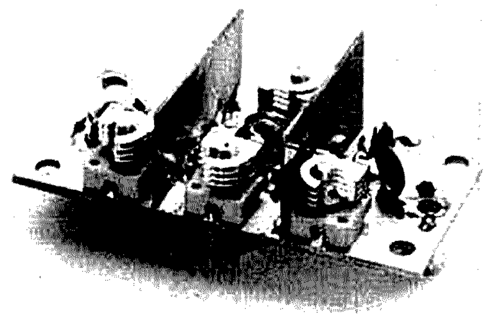


fig. 6. Full-sized printed-circuit board for the uhf preamplifier. Component layout is shown in fig. 5.



Uhf preamplifier for 432 and 450 MHz uses two low-noise, grounded-gate fet stages.

6. Form the fet leads, and trim the leads to within 1/16 inch (6.5mm) from edge of transistor case. Slide the drain lead under the shield notch and seat the transistor in the hole in the circuit board. Carefully tack solder the leads close to the case, but do it quickly to avoid overheating the fet.

7. On the fiberglass side of board install the two rf chokes between C8, C10 and C12, leaving room between RFC2 and J2. Install R1 and R2 from C7 and C9 to the board. Then install diode CR1 with the cathode to feedthrough capacitor C11 and anode to ground.

Diode CR1 is used to protect the fets from reverse transients on the power wiring. However, diodes should also be installed across relay coils and other inductive components sharing the B+ line to prevent fet damage from voltage spikes which are generated by collapsing magnetic fields.

The preamp may be aligned either in conjunction with a receiver or with a 50-ohm load and an rf detector and sensitive vtvm. The best signal source is a signal generator with a variable attenu-

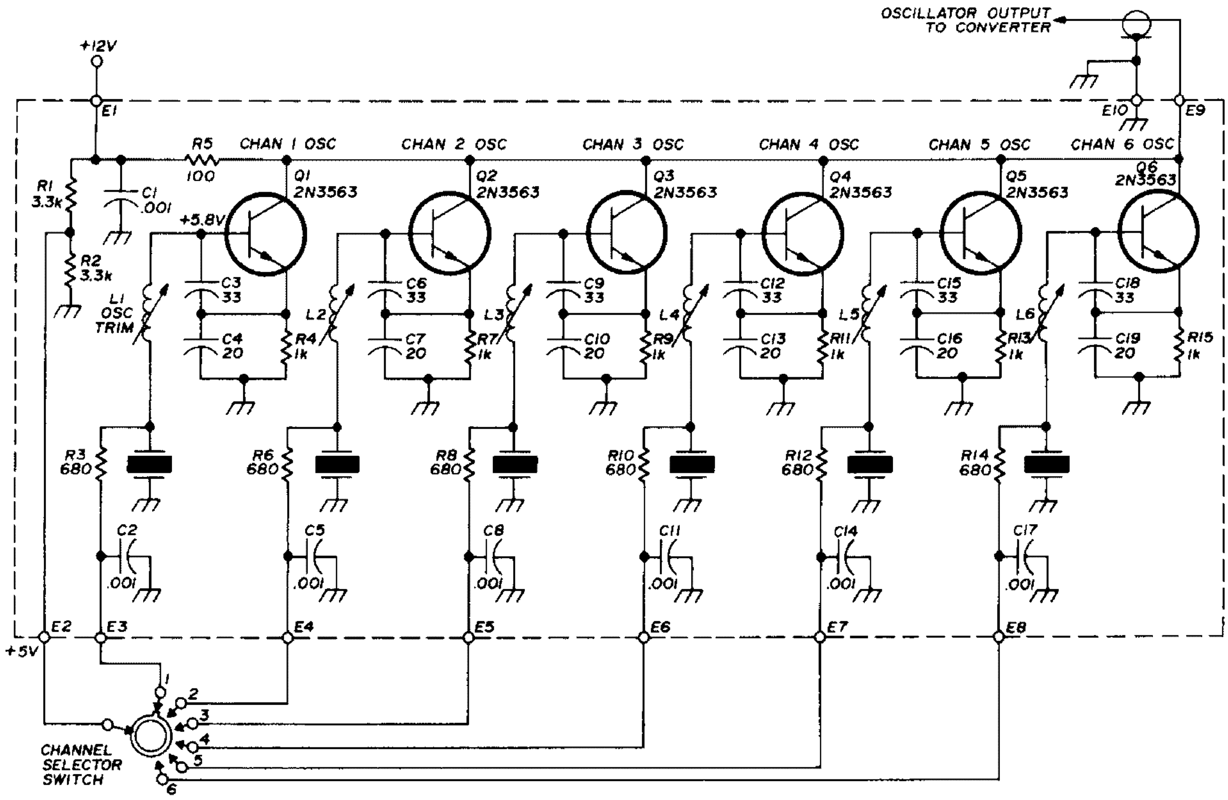


fig. 7. Schematic diagram of the six-channel oscillator for use with the uhf converter. Inductors L1-L6 consist of 11½ turns no. 26 (0.4mm) wire on ¼" (6.5mm) diameter forms with carbonyl-J slugs.

ator; however, a strong transmitted signal may also be used. An insulated tuning tool should be used for tuning.

Capacitor C2 will tune broadly while C3 and C5 tune sharply. In addition to resonating L3 to the desired frequency, C4 also establishes loading of the first stage to some extent. If there is a tendency for the preamp to oscillate, either the load on J2 is greater than 50 ohms or C4 needs to be increased to place a heavier load on Q1. Capacitors C3 and C5 should always be tuned last, because they tune fairly sharply. If one of the capacitors peaks at maximum capacitance, the coil should be enlarged (and vice-versa). Some allowance can be made when the preamp is first built if the frequency is far removed from the 440-MHz design center.

multichannel operation

If a multichannel oscillator, such as the one shown in **fig. 7**, is used in place of the built-in oscillator on the converter board, the components marked with an asterisk in **fig. 1** should be removed from the converter. The output cable from the multichannel oscillator should be connected to L4 (Q4 collector connection in **fig. 1**), and the shield should be connected to a nearby ground. The six-channel adapter (or as many as you want to use) consists of independent oscillators similar to the one on the converter board. Five volts are applied to individual oscillators to turn them on at appropriate times. DC switching can be done either with an electromechanical switch or with a scanner adapter.[3]

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2. Bruce Clark, K6JYO, "A Stable, Variable-Output Weak-Signal Source," *ham radio*, September, 1971, page 36.
3. Jerry Vogt, WA2GCF, "Improved Channel Scanner, for VHF FM," *ham radio*, November, 1974, page 26.

parabolic reflector gain

Graphical method for determining gain and beamwidth in terms of frequency for dish diameters up to 100 feet

Walter E. Pfiester, Jr., W2TQK, 1 Skadden Terrace, Tully, New York 13159

The parabolic reflector, used in conjunction with an efficient feed, is probably the best antenna available today for use at frequencies over 300 MHz. Many articles have been written on the construction[1-4] and feed systems[5,6] for proper illumination of the dish; however, few have examined the gain from a mathematical point of view. The interrelationships of antenna gain, beamwidth, and dish size are discussed in this article and represented graphically on an easy-to-use linearized nomogram.

antenna gain and beamwidth

The equation usually used to describe the available gain from a parabolic reflector, well known to all microwave engineers, is:[7]

$$G_a = 7.5 + 20 \log D + 20 \log F \quad (1)$$

where G_a = gain above isotropic (dB)
D = diameter of the dish (feet)
F = frequency (GHz)

When the diameter of the dish is given in metric terms, the gain is given by:

$$G_a = 17.82 + 20 \log D + 20 \log F \quad (2)$$

where D is the diameter of dish (meters), the other terms the same as in **eq. 1**.

Interrelated with the gain equation is the beamwidth of the main lobe, which is approximated by:

$$\emptyset = \frac{70}{D \times F} \qquad (3)$$

where $\emptyset$ = total half power or 3 dB bandwidth (degrees)
D = diameter of the dish (feet)
F = frequency (GHz)

Again, when the diameter of the dish is given in metric terms, the beamwidth of the main lobe is approximated by:

$$\emptyset = \frac{21.3}{D \times F} \qquad (4)$$

where D is the diameter of the dish (meters), the other terms the same as in **eq**. 3.

These equations assume that the antenna has 55 percent efficiency and tapered illumination such that the illumination on the outer rim of the dish is 10 dB down from the illumination at its center or bore site. For example, what is the theoretical gain and half-power beamwidth of a 30-foot parabolic reflector on 432 MHz?* From **eq. 1** the gain is

$$\begin{aligned} G_a &= 7.5 + 20 \log 30 + 20 \log 0.432 \\ &= 7.5 + 20(1.48) + 20(9.64 - 10) \\ &= 7.5 + 29.54 - 7.29 = 29.75 \text{ dBi} \end{aligned}$$

The half-power beamwidth from **eq. 3** is

$$\emptyset = \frac{70}{30 \times 0.432} = 5.4 \text{ degrees}$$

These equations are obviously unwieldy, even with slide rules built specifically for this purpose,[8-10] especially in those cases where you are trying to compare the various factors which are involved. **Fig. 1** is a graphical solution to these equations. The diameter of the dish, from 1 foot (30cm) to 100 feet (30.5 meters), is plotted along the horizontal axis of the graph. Along the vertical axis, on the left-hand side, is the gain of the antenna in dBi (for gain above a reference dipole, subtract 2.15 dB from the value shown on the graph). Values are represented from zero to +70 dBi.

The right-hand vertical axis represents the antenna beamwidth (main lobe) in degrees at the half-power points. Values from 128 to 0.06 degrees are shown. The family of curves shown by the diagonal lines cover frequencies over the range from 100 MHz to 32 GHz. The graph is constructed so that values not shown are easily interpolated and may be drawn on the graph with a straight edge.

using the nomograph

Assuming a recent acquisition of a surplus 6-foot (1.8m) dish, what can you expect from this dish at 432 and 1296 MHz? Draw a vertical line from the 6-foot (1.8m) point of the horizontal axis to the 432 and 1296 MHz curves (interpolated on **fig. 1**). Draw horizontal lines from these points to both the left and right axis. At 432 MHz, the nomograph shows that it's possible to get 15.5 dBi gain with a beamwidth of 30 degrees. At 1296 MHz, this same dish will deliver 24.5 dBi gain with a beamwidth of 9 degrees.

*Using metric dimensions, what is the gain and half-power beamwidth of a 9.14-meter (30-foot) parabolic reflector at 432 MHz? From **eq. 2** the gain is

$$\begin{aligned} G_a &= 17.82 + 20 \log 9.14 + 20 \log 0.432 \\ &= 17.82 + 20(0.96) + 20(9.64 - 10) \\ &= 17.82 + 19.22 - 7.29 = 29.75 \text{ dBi} \end{aligned}$$

From **eq. 4** the half-power beamwidth is

$$\emptyset = \frac{21.3}{9.14 \times 0.432} = 5.4 \text{ degrees}$$

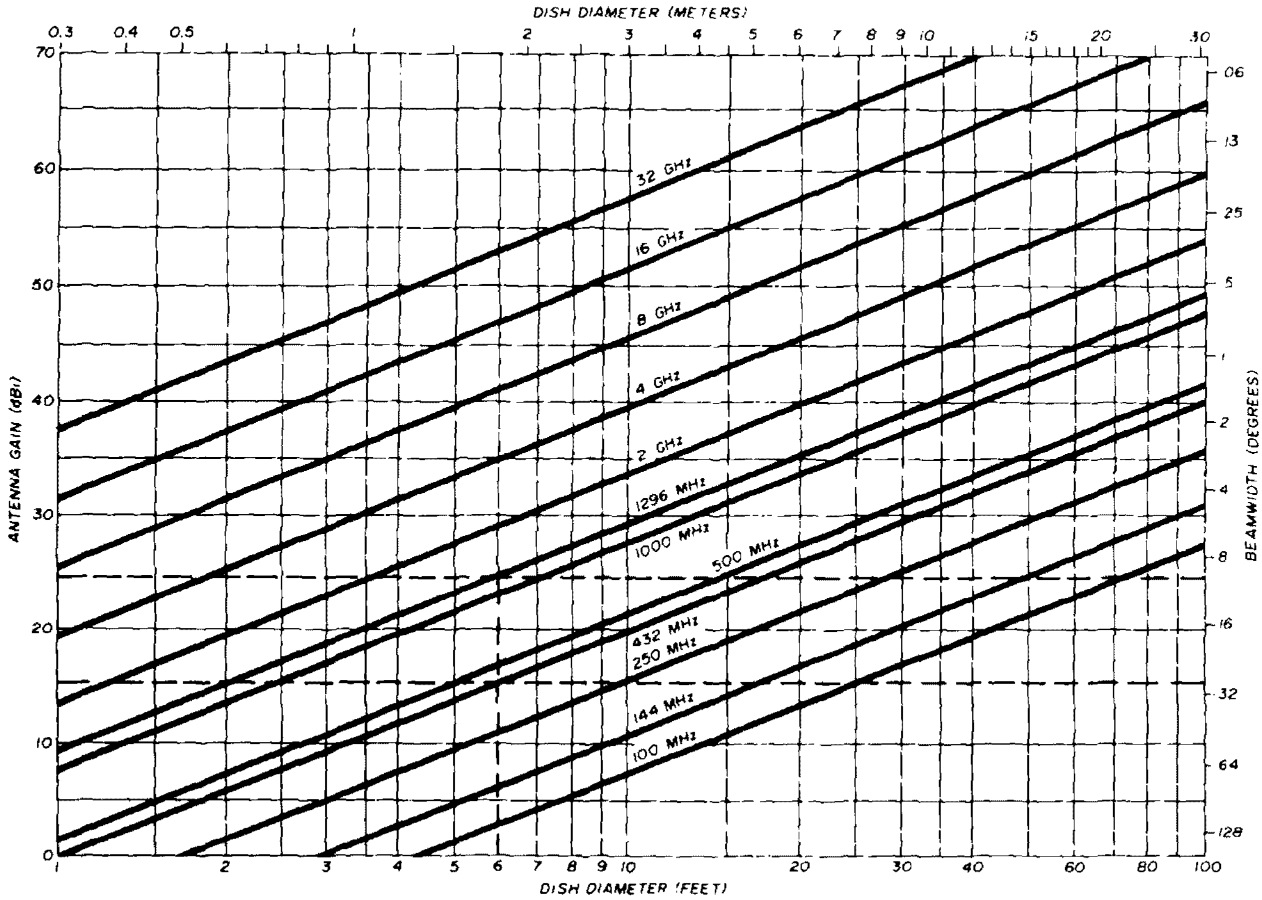

fig. 1. Antenna gain and beamwidth versus dish diameter. Dashed lines are examples discussed in the text.

These gain figures are practical and reliable, having been attained by other than professionals in the microwave field. Through judicious use of this chart tradeoffs may be made easily among beamwidth, gain, and size, with real-world values resulting. I'd appreciate hearing from those having any trouble with the use of this chart and would appreciate comments on its use.

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6. Dolph Vilardi, WA2VTR, "Simple and Efficient Feed For Parabolic Antennas," *QST,* March, 1973, page 42.

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ham radio

1975 ham radio sweepstakes winners

WB6QVW is the grand prize winner — eight others win either Atlas transceivers or Icom IC-230s

Skip Tenney, W1NLB, Publisher, Ham Radio Magazine

The sixth annual *Ham Radio* Sweepstakes is over. After many busy weeks, both at our local post office and in our Greenville office, the many hours of opening and reading your sweepstakes entries are finally over.

Much as we like all of the Sweepstakes excitement it was quite a relief when May first finally rolled around and we knew that it was over for another year. There was an excellent crop of entries this year, and they were matched with more equipment prizes than we have ever offered before.

This year's prizes centered around two basic choices. Four winners received their choice of Atlas Radio's exciting 210 or 215 solid-state single-sideband transceivers while another four

Fred Moller, WN1USO, Ham Radio Advertising Manager, contemplates the combination of an Atlas 210 and a General Class license.

Ham Radio receptionist Rose Jenkins picks the lucky winner, WB6QVW.

winning tickets entitled their owners to the very much sought after Icom IC-230 synthesized two-meter transceivers.

The luckiest guy of all was Andy Ellis, WB6QVW, who won our grand prize. He received both an Atlas and an Icom transceiver. Andy can now boast of having the very latest, both in single-sideband and in fm equipment. He is particularly excited about his prizes and plans to use his new Atlas 210 as a downlink receiver for Oscar work and the IC-230 to compliment the Icom IC-21 he already owns to give himself both mobile and base capability on two-meter fm.

Bob Hueberger, W2NWE; George Pastilla, W6RRC; Con Weigand, WA8SCA; and David Kochendarfer, K4DC were the fortunate winners of the new ultra-compact Atlas transceivers. These 200-watt rigs feature broadband tuning and a solid signal on either ssb or CW. They operate on five bands; the model 210 handles 80, 40, 20, 15 and 10 meters while the model 215 covers 160 meters in place of 10 meters.

These rigs feature the smallest physical size and one of the hottest receivers in amateur radio today. Operating from 12 Vdc they certainly represent one of the outstanding mobile high frequency transceivers ever offered to the radio amateur. These rigs turned the eyes of all the radio amateurs on our staff while they were here waiting for their new owners.

Icom's fabulous IC-230 PLL synthesized two-meter transceivers were the perfect prizes for Mike Bardon, WA7ZGI; Earl Cunningham, W5RTQ; Ira Cohn, K2CTK; and Charlie Spencer, K4RXX.

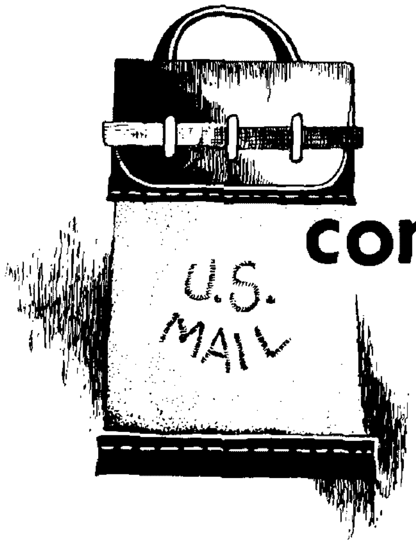
This is the rig that has really been setting the pace in two-meter circles. Covering all of the standard 30-kHz fm repeater pairs with its own easy-to-use, built-in synthesizer, these rigs can also cover a number of the new 15-kHz channels with merely the addition of appropriate crystals.

The IC-230 also boasts a very sensitive receiver front end which will match the performance of virtually any amateur fm installation. All of this convenience and performance is put in a compact package no larger than most of the crystal-type rigs now on the market.

Ham Radio publisher, W1NLB, checks in on a local repeater with one of the Icom IC-230s given away in this year's sweepstakes.

We had a lot of fun bringing you this year's Sweepstakes and a large vote of thanks is due to the many thousands of you who entered. Let's hope *you* win next year.

ham radio



comments

wind loading on antennas

Dear HR:

The recent article by John Nagle, K4KJ,[1] is is very excellent educational article on the wind loading of *towers* and gives data and recommended design coefficients which I have been searching for. However, on the subject of wind loading *calculations for antennas*, I have a question. Author Nagle would merely choose the largest calculated area looking end-on to the boom or end-on to the elements. Unlike the case for the guyed tower with the wind blowing at an angle with respect to the guy wires, he ignores this possibility when calculating the wind loading of beam antennas. I should also mention that he is not the only author who has used this simplified approach.[2]

Let's take a specific case: a five-element, 20-meter Yagi. To simplify the arithmetic, assume a 3-inch (7.6cm) diameter boom 48-feet (14.6m) long with 1-inch (2.5cm) diameter elements, all 33 feet (10.1m) long. Using Nagle's approach for the boom case (wind blowing perpendicular to the boom), the antenna area is

$$3/12 \times 48 \times 0.66 = 7.92 \text{ square feet } (0.736\text{m}^2)$$

The author would choose 9.075 square feet (0.843m^2) as the projected area to be used in calculating the wind loading. However, the wind may not be so accommodating. Suppose that the wind is blowing at 45° with respect to the boom and elements. This would result in a force equivalent to the vector sum of *both* projected areas:

$$(7.92 \sin 45^\circ) + (9.075 \cos 45^\circ) =$$
$$(7.92 \times 0.707) + (9.075 \times 0.707) =$$
$$12.015 \text{ square feet } (1.117\text{m}^2)$$

Since designs should always be for the worst case, I maintain that this approach should be used for antenna wind-loading calculations. As the referenced *QST* article points out, gusty winds can often act in concert with the natural oscillation of the tower/antenna system to produce deflections which are many times those expected of a steady wind. If we kid ourselves about how much antenna area we have on the top of our towers, the result could be painful. I, for one, am accepting no more manufacturers' ratings in this category without first performing my own calculations.

I have heard it said that the case I pose results in an overstated wind load

1. John Nagle, K4KJ, "How to Calculate Wind Loading on Towers and Antenna Structures," *ham radio*, August, 1974, page 16.
2. Richard Lodwig, W3GNK, "Wind Force on a Yagi Antenna," *QST*, July, 1974, page 46.

because the boom would "shadow" the elements behind it. A year of observing wind velocity and direction, with sensors located at the top of my tower, has shown that neither wind velocity nor direction shows any consistency for durations longer than 3 or 4 seconds. This is anything but laminar flow and I would not count on any shielding effect of the boom in lessening wind load on the antenna elements.

Forrest E. Gehrke, K2BT
Mt. Lakes, New Jersey

Mr. Gehrke is, of course, quite correct in his statement that one should take the vector sum of all forces acting on the antenna just as for the tower case. For the boom and element diameters assumed by reader Gehrke, the maximum effective area is 12.045 square feet (1.119m²) and occurs at an angle of 41.11 degrees.

Using differential calculus and the terminology given in **fig. 1**, *the wind angle that sees the maximum area, and hence will develop the maximum force, is given by*

$$\theta = \arctan (A_b/A_e)$$

The maximum effective area, A_{eff}, is given by the Pythagorean theorem as

$$A_{eff} = \sqrt{A_b^2 + A_e^2}$$

The reason Mr. Gehrke came as close to the maximum force as he did in assuming a wind angle of 45 degrees is because his element and boom areas are almost equal and the actual maximum wind force angle of 41.11 degrees is close to the 45 degree angle he assumed. In the more general case, however, the two areas will not be equal and the equations given above should be used.

If one does not have an inclination for mathematics and wants an easy, conservative approach, the boom and element areas can be added arithmetically. Since the arithmetical sum will always be greater than the vector sum, this will lead to a conservative design.

As to K2BT's doubts concerning the shadowing effect of the larger diameter boom on the smaller diameter elements, I concur. Even with laminar air flow (non-turbulent) where some shadowing may exist, I doubt that the shadowing

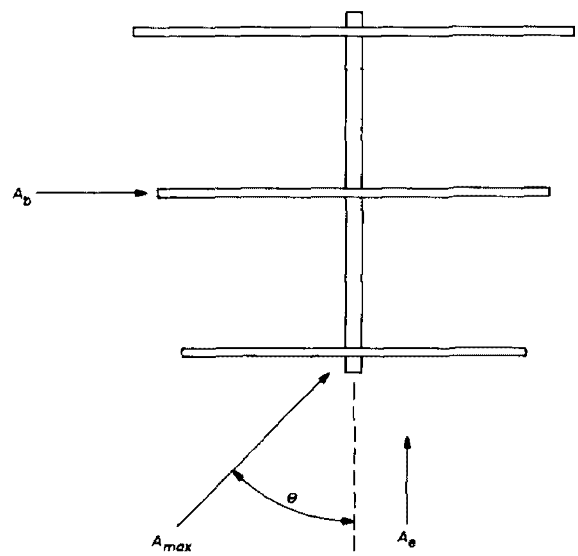


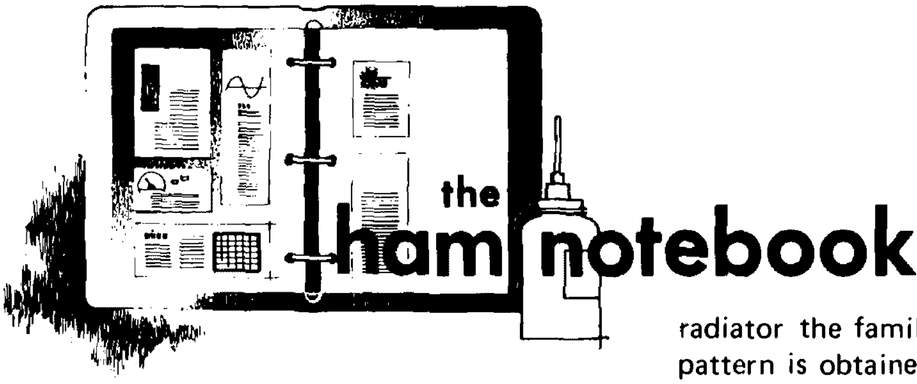
fig. 1. The angle at which the maximum area is presented to the wind is given by θ = arctan A_b/A_e.

extends further than about one boom diameter behind (on the lee side) of the boom. When dealing with element lengths that may extend 5, 10 or even 20 feet (1.5, 3 or 6m) on each side of the boom, the shadowing, if it does exist, will be negligible and can be neglected for all practical purposes.

Regarding Mr. Gehrke's statement that he is no longer accepting manufacturers' antenna wind-loading ratings but will make his own calculations, I would like to quote one of Murphy's lesser known laws:

"Manufacturer's spec sheets will be incorrect by a factor of 0.5 or 2.0, depending upon which multiplier gives the most optimistic value. For salesman's claims, these factors will be 0.1 or 10.0."

John J. Nagle, K4KJ
Herndon, Virginia



432-MHz OSCAR antenna

The antenna described here is a simple approach for 432-MHz stations who want to operate on the OSCAR 7 uplink and have high output power (100 watts or greater). It is similar to the slanted ground-plane antenna discussed by K4GSX[1] but does not use complex impedance matching networks or tuned lines.

The antenna, shown in **fig. 1**, consists of a ground plane and folded monopole radiator. By tilting the angle of the radiator the familiar distorted radiation pattern is obtained. The use of a folded monopole provides protection from electrical discharge and also results in a very low vswr – typically 1.2:1 – without any special matching schemes.

Construction details for the slanted 432-MHz monopole are shown in **fig. 1**. The ground plane is easily made from a piece of sheet aluminum (such as the Reynolds *Do-It-Yourself* material found in many hardware stores). A type-N connector mounted at the center of the ground plane provides the connection to the radiator and serves as the mounting terminal for the folded monopole. One side of the monopole is soldered to the center pin of the connector and the other side is soldered to a grounding lug which is held in place by one of the connector-mounting screws. A dab of RTV (or similar sealant) over the connector provides weatherproofing and completes the construction.

While this simple antenna is not comparable to a fully steerable array, it provides good omni-directional coverage whenever the satellite is above 10 to 20

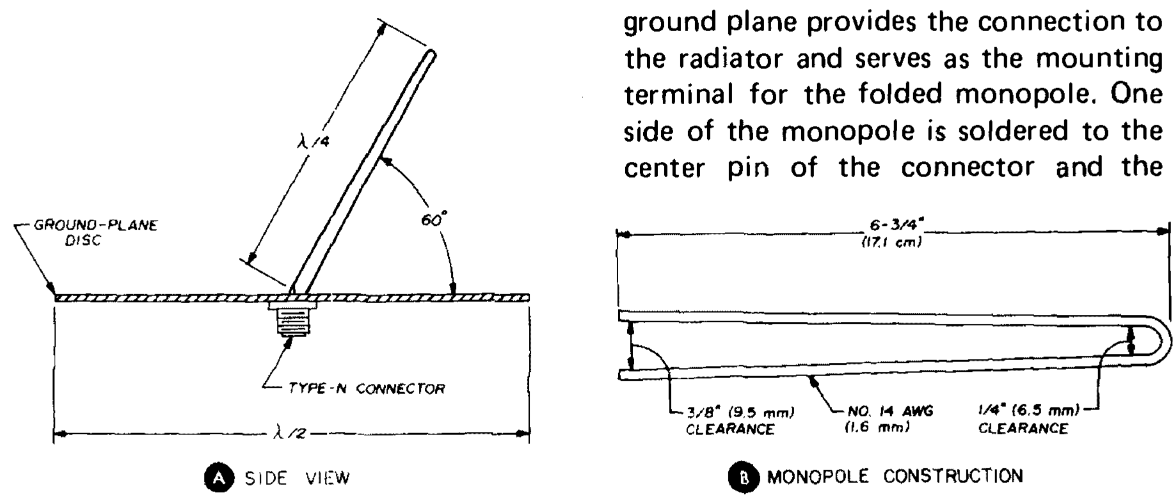


fig. 1. Ground-plane antenna with slanted folded-monopole radiator for use on the OSCAR 7 432-MHz uplink. Ground-plane disc is 13 to 15 inches (33 to 38 cm) in diameter (not critical), 1/16" (1.5mm) or thicker aluminum. Construction of the monopole radiator is shown at B.

1. Dale Covington, K4GSX, "Simple Antennas for Satellite Communications," *ham radio*, May, 1974, page 24.

degrees elevation. Below that elevation, the normal high-gain 432-MHz antenna can be switched in for the real DX.

Joe Reisert, W1JAA

audio transducer

The audio transducer shown in the photograph (fig. 2) with all the other gear is a modified permanent-magnet speaker with its voice coil removed and its diaphragm intact. It is a tactile aid. Mine was made from an old PA system, but it can be made from any loudspeaker. The core is magnetic and vibrates strongly at a low audio frequency.

fig. 2. Gayle Sabonaitis, WA1OPN, can copy CW up to 25 wpm using the audio transducer she is using in this photograph.

Building the transducer is quite simple — all that is needed is a discarded loudspeaker. When the coil is removed, it is advisable to enclose it in a metal box so it will not get crushed (the plastic cone is easily broken). It is a good idea to use an 8-ohm speaker as deaf persons respond best to low-impedance speakers because the vibrations are stronger.

Gayle Sabonaitis, WA1OPN, a student at the Perkins School for the Blind in Watertown, Massachusetts, is both blind and deaf and has been licensed for three years. She is presently preparing to take the Advanced class exam.

I have used an audio transducer for five years and find it much easier than headphones. I do not have any hearing left — it all disappeared a few years ago and headphones can be very confusing — it is difficult to copy CW with only a little hearing left. I can copy CW at 25 wpm and send at 18 wpm. These speeds may appear to be an overstatement when said by a deaf amateur who has only been on the air three years, but they are not an exaggeration and are possible with the help of an audio transducer. However, I usually copy about 18 wpm when there is too much interference. Some deaf-blind hams can copy faster than this because they have had much more practice.

Gayle Sabonaitis, WA1OPN

gated oscillator

The circuit shown in fig. 3 is a versatile, minimum-cost oscillator using a minimum of components. It can be made to operate from less than 100 kHz to over 80 MHz by selecting a 74S00 for any operating frequency above 20 MHz. The output is a square wave swinging between about 0.4 volt and 4.5 volts. The oscillator can be started and stopped by switching the oscillator gating input between zero and 5 volts. If this feature is not desired, the NAND gates may be replaced by three sections of a 7404 hex inverter, resulting in a package count savings of one fourth. All external components are the same as those shown.

Doug Schmieskors, WB9KEY

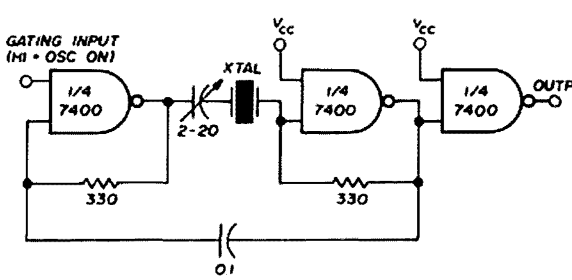
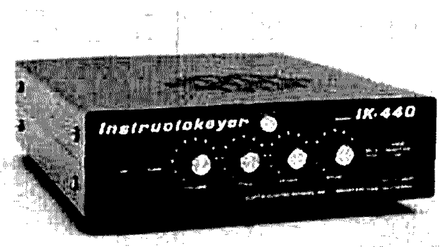


fig. 3. Simple gated oscillator circuit can be used from 100 kHz to 80 MHz. For use above 20 MHz, select higher speed 74S00 ICs.

new products

code-teaching keyer



Code practice has suddenly become more efficient and interesting because of a new development from Curtis Electro Devices. Their new IK-440 Instructokeyer provides an infinite variety of code groups allowing unlimited practice for higher proficiency. In addition, the IK-440 is a state-of-the-art keyer which uses the new 8043 keyer-on-a-chip IC. It uses no motors or tapes.

The all solid-state unit sends random groups of Morse letters, numbers, punctuation and word spaces in an ever changing sequence which never exactly repeats. Code speed is adjustable from 4 to 50 wpm. Code groups are of varying lengths but average five characters per group. A rear panel switch selects alphabet only or full alphanumeric.

The keyer portion of the IK-440 provides self-completing dots, dashes and spaces, instant start, dot memory, weight control, iambic mode and built in sidetone. Front panel controls are provided for volume, pitch, weight, speed, on-off, tune, self-test and "keyer-code practice." The IK-440 will key ±300 Vdc at 200 mA in either the keyer or code practice mode. Output switching is solid state. Price of the IK-440 is $224.95.

For further information, contact Curtis Electro Devices, Inc., Box 4090, Mountain View, California 94040, or use *check-off* on page 94.

500 MHz prescaler

Levy Associates has broken the frequency-measurement barrier with a prescaler that will count to 500 MHz. Requiring only 150 millivolts to operate, this prescaler is guaranteed to respond to 500 MHz minimum, and will typically operate to 550 MHz. Division by ten and one hundred is provided, and the TTL compatible outputs provide adequate drive for all commercial and home-built counters.

The combination of high input impedance (500 ohms) and high sensitivity makes possible easy measurement of low-power transmitters. The frequency of a one-watt transmitter can be measured at four to six feet using only a ¼-wave whip antenna to drive the prescaler. Overload protection is provided up to 2 volts input.

The prescaler is complete including 117-Vac power supply on a 3x4-inch (7.6x10.1-cm) circuit board, small enough to fit inside many counters. Available from stock as a kit, with all

More Details? CHECK—OFF Page 94

parts, drilled circuit board and instructions, $89.00; or completely assembled and tested, $109.00 (plus $.85 postage and California sales tax if applicable). For more information write to Levy Associates, Post Office Box 961R, Temple City, California 91780.

ameco equipment

The popular line of Ameco amateur equipment is now available directly from the manufacturer. Included in the line are lowpass transmitting filters, highpass TV filters, all-band preamplifiers, a standing wave bridge and bridge indicator unit, code practice oscillators, and a novice CW transmitter kit. Also available are heavy duty 572B/T160L power triodes (replace 811As), type-UHF rf connectors and economy slide switches. For a copy of their latest catalog, write to Ameco Equipment Company, 314 Hillside Avenue, Williston Park, New York 11596, or use *check-off* on page 94.

MFJ catalog

The new MFJ catalog of amateur radio equipment describes many *new* products, including CW and ssb audio filters, electronic keyers, frequency standards, audio amplifiers, active filters, printed-circuit boards and electronic components. Copies of the catalog and additional information are available from MFJ Enterprises, Post Office Box 494, Mississippi State, Mississippi 39762, or use *check-off* on page 94.

short circuit

The new Jackson Brothers high-precision trimmer capacitor described on page 65 of the March, 1975, issue of *ham radio* is known in this country as the *N.R.P.* capacitor. *Trimline*, the name used in England, cannot be used in this country as it is a trade name owned by another firm.

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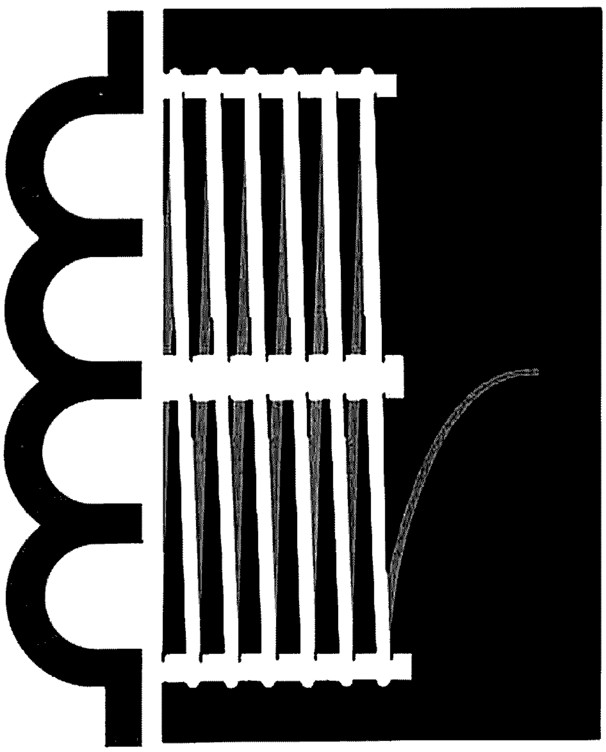
AUGUST 1975

500-WATT POWER AMPLIFIER FOR 160 METERS



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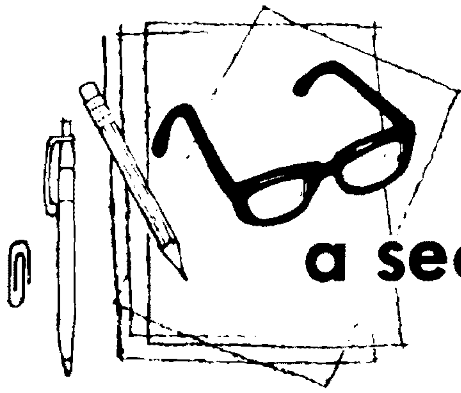

a second look by jim fisk

It doesn't seem possible, but it was only ten years ago *that the price of field*-effect transistors finally dropped to the point where it became possible to use them in amateur radio projects. Those early fets, which were limited to audio frequencies, were soon replaced by devices which worked well into the high-frequency range, and later, to vhf and uhf. Now, advances in gallium arsenide (GaAs) technology have produced field-effect transistors which exhibit an *available* gain of 18 dB and minimum noise figure of 4 dB at, get this, 10,000 MHz.

If this sounds like science fiction, consider the fact that these 10-GHz fets are only the tip of a much larger iceberg – by the end of the year it's expected that advanced devices will move into the 15-GHz region, and in several years, possibly to 25 or 30 GHz. Although practical circuit experience is still somewhat limited, many researchers feel that these new devices combine the best characteristics of Schottky and tunnel diodes, but without the isolation problems inherent in two-terminal devices.

Although these fets are usually thought of as small-signal devices, cellular GaAs fets have been combined to form very efficient linear amplifiers which rival the best bipolar linears down to about 4 GHz (gain is so high at lower frequencies it's practically impossible to build unconditionally stable amplifiers).

Most of the research so far has been devoted to single-gate fets, but dual-gate structures have been built and they show great promise. Their noise figure is a bit higher than the single-gate fet, but the dual-gate version has higher gain. Furthermore, the dual-gate GaAs fet has a large gain modulation range so it's suitable for use with agc.

In addition to their use as linear amplifiers, these fets are finding applications as oscillators and mixers for integrated microwave receiver front ends. Since the GaAs fet exhibits a large dynamic range with conversion gain, it may eventually replace diode mixers at microwave frequencies. As an example of an fet mixer, scientists at Raytheon recently showed an 8-GHz GaAs fet mixer with a 7.8 dB noise figure and third-order intercept point at +18 dBm. Low-level tunnel-diode mixers for this same frequency have a noise figure which is about 1 dB lower, but gain is also lower and the third-order intercept point is only +5 dBm. However, other researchers have shown that the noise figure of the fets can be reduced considerably by cooling and predict that 1 to 2 dB noise figures will eventually be possible at 10 GHz.

So we come *full circle* . . . ten years ago a 1 or 2 dB noise figure at 144 MHz was just barely possible, but only *if* you were willing to use a complex parametric *amplifier*.

Jim Fisk, W1DTY
editor-in-chief

presstop

REPEATER LINKING (Docket 20073) okayed by the FCC, will be permitted after July 11th. Limitations are that linked repeaters must operate on the same band (cross-banding linked repeaters will be covered later with Docket 20113), licensee of each repeater in a linked system must submit a new system network diagram showing all stations in the system.

Repeater Automatic Control (Docket 20112) also approved, requires that repeaters operated without a control operator must record their transmissions for later review, incorporate procedures to shut the system down for "malfunctions or improprieties." No modification of existing repeater licenses is required by the terms of the Report and Order, which goes into effect July 28th.

"Closed Repeaters" (defined as "repeaters used only by persons specifically authorized by the control operator") are exempted from the requirements for recording and review — such systems typically require some form of coded access.

HIRAN DOCKET is definitely warming up as FCC's Office of Chief Engineer has asked Amateur And Citizen's Division for its list of coast area frequency coordinators. Lab and/or field testing to determine interference potential is a strong possibility, and if such tests get complicated they could further delay any decision on whether HIRAN will end up in the 420-450 MHz Amateur band.

WARC 1979's last open slot, chairmanship of 1296 MHz and up Task Force, has been filled by Chuck Dorian, W3JPT. Initial work of some WARC groups has begun with mailings and telephone discussions.

MID-CONTINENT AMSAT NET is moving from 3850 to 7280 kHz for its regular 9:00 PM Tuesday night (0200Z Wednesday morning) sessions, at least for the summer. This should reduce QRN and propagation problems, may encourage some additional participation from AMSAT members not on 75 meters.

Another Mode Jump occurred in OSCAR 7 on June 22nd. Any listener who heard the satellite go from Mode A to Mode B or simply didn't find it on the scheduled mode as expected could help AMSAT by dropping a note to Box 27, Washington, D.C. 20044.

New Satellite Amateur Band proposed by Bob Haviland, W4MB, head of the WARC '79 Task Force on 27-1296 MHz amateur allocations. A "propagation optimum" exists in the 900-MHz region, so Bob proposes a secondary amateur allocation for space communications only in the 33cm area. He also proposed that satellite allocations in other amateur bands be broadened.

Another AMSAT Project Leader is sought, this time to oversee design and construction of low cost 2- or 10-to-BC band converters (with bfo) for use in schools with AMSAT's educational program.

SEVERE RESTRICTIONS ON TRANSMITTERS proposed for Texans. A set of regulations restricting all rf electromagnetic radiation has been written up by the Texas State Department of Health Division of Occupational Health and Radiation Control. They would effectively prohibit hand-held or even mobile transmitter use in the state because of their unreasonably low levels of permitted exposure. In addition, the proposed regulations would severely limit the power output of fixed stations — and would, of course, cover public service and broadcast as well as Amateur and CB.

Whether A State even has the power to make such restrictions is questionable — the Communications Act of 1934 gives the Federal government jurisdiction over radio communications — but if the proposal should be adopted anyway, a lengthy (and expensive to the taxpayer) court battle would undoubtedly result.

500-watt power amplifier for 160 meters

Modifying the Heath HA-14 or SB-200 for top-band operation

Albert Segen, W2BP, 101 Collins Avenue, Pleasantville, New Jersey

With the recent decrease in sunspot activity and fewer DX openings on 10 and 15 meters, interest in the 160-meter band operations has increased markedly. And, in the future, with Loran A vacating the band, we can expect full use of 160 meters with restoration of full power.

I first got on 160 in late 1968, completed my 160-meter WAC in 1973 and my 160-meter country total now stands at 74 with nearly half on ssb. The 160-meter band is now my primary amateur interest and in April, 1972, I went on a one-operator, seven-country, 160-meter DXpedition into the eastern Caribbean.[1]

Amateur stations in the states and provinces along the Atlantic and Pacific coasts are generally restricted to 100 watts input power at night and 500 watts during the day. The idea of 500 watts of daytime power is very appealing, particularly for use during those short periods around sunrise and sunset when long-range DX contacts can be made on occasion.

Sometime ago I picked up a Heath HA-14 linear rf power amplifier for use on 10 through 80 meters. The HA-14, known as the "Kompact KW," was originally designed for mobile use although separate ac and dc power supplies were available. It appeared that the HA-14 could be modified for operation on 160 meters, so another was obtained. The HA-14 is no longer in production but units invariably turn up at amateur flea markets and auctions. The current pro-

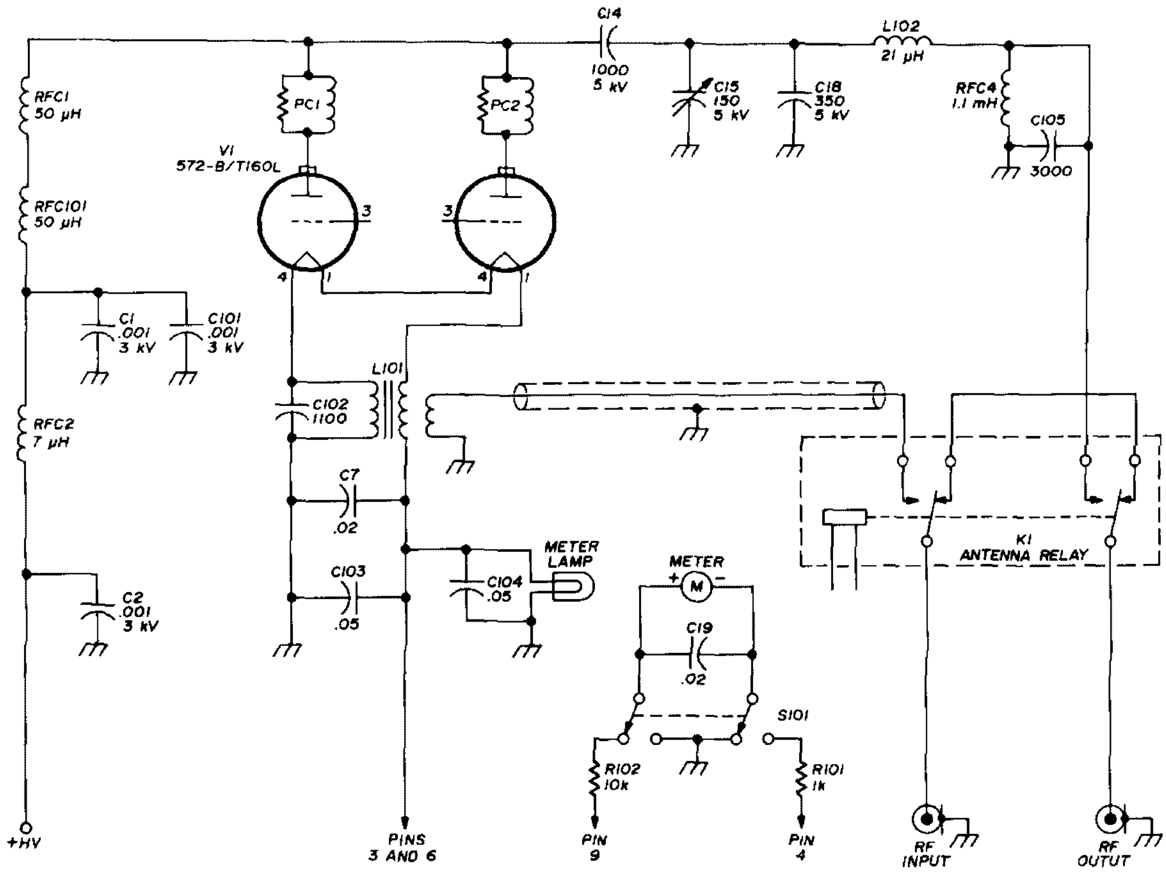

fig. 1. Schematic of the Heath HA-14 modified for operation on 160 meters. Components with 100-series part number (R101, C101, etc.), are added.

duction Heath SB-200 is much like the HA-14 rf circuit; modifications unique to the SB-200 will be treated later in this article.

The modifications described are for a 1 kW ssb amplifier operating over the range from 1800 to 2000 kHz. Some performance will be compromised when the amplifier is used at a CW level of 500 watts input.

design factors

Before modifying a proven unit like the HA-14, I felt it was appropriate to go through some of the basic design calculations for the output pi-network. Using a plate supply voltage of 2100 volts and an input of 1 kW peak ssb in class-B linear operation, the plate load impedance was calculated to be about 2800 ohms. This plate impedance requires a pi inductance of about 21 μH and a tuning capacitance around 440 pF.[2] Happily, these values could be accommodated in the space available within the HA-14.

With the HA-14 operating at 500 watts input on CW, the plate load impedance is essentially doubled, requiring an inductance far larger than the room available in the HA-14. Therefore, using component values calculated for 1 kW ssb input, amplifier efficiency is somewhat lower at 500 watts CW but there still is a worthwhile power increase.

modifications

A schematic diagram for the 160-meter HA-14 power amplifier is shown in **fig. 1**. As there are no changes required in the grid circuit, that portion of the amplifier is not included in the diagram. The values of the original plate choke, RFC1, and bypass capacitor, C1, are not high enough for operation on 160 meters. The addition of choke

RFC101 in series with RFC1 and capacitor C101 in parallel with C1 solve this problem. The location of these parts is shown in fig. 2.

All components in the plate circuit tank compartment are removed except the variable capacitor and the wiring harness. I chose not to destroy the original coil and was able, with care, to disassemble the multideck bandswitch by loosening all the assembly screws. This allowed the original coil to be removed without cutting any coil connections to the deck switch.

The output capacitor, C105, consists of three capacitors in parallel which are mounted on the rear wall of the compartment. One is a husky 2000 pF mica while the others are 500 pF postage-stamp micas (see fig. 3). Do not skimp on the quality or size of these capacitors as considerable rf current flows at this point.

The pi network inductance, L102, has 37¼ turns, the quarter turn used to provide mechanical support as can be seen in fig. 3. When initially cut from a B&W 3026 air inductor (2 inch [51 mm] diameter, 8 turns per inch, [3.1 turns per cm], 10 inches [25cm] long), 39 turns were used. Almost two turns were unwound at the antenna end of the coil to permit the use of the polystyrene support rods as short standoff insulators for support at the left side of the compartment. A ceramic standoff insulator at the antenna end provides further mechanical support.

The 350 pF capacitor (C18 in the original circuit) is used as part of the plate tuning circuit and is mounted on a piece of aluminum strip along with the coupling capacitor, C14. Capacitor C18 is installed on the right side of the compartment and acts as a support for the hot end of inductor L102.

All the input coils and capacitors are removed at the same time the original tank circuit is removed. I was not successful in designing an input pi-network with essentially flat vswr over the range of 1800 to 2000 kHz without some sort of retuning at each end of the band. The circuit I finally worked out proved to be simple, effective and used a minimum of parts – inductive coupling of the exciter into the bifilar filament choke. The bifilar choke remains in place as in the original HA-14 except that a third winding is carefully wound onto the bottom end (see fig. 4).

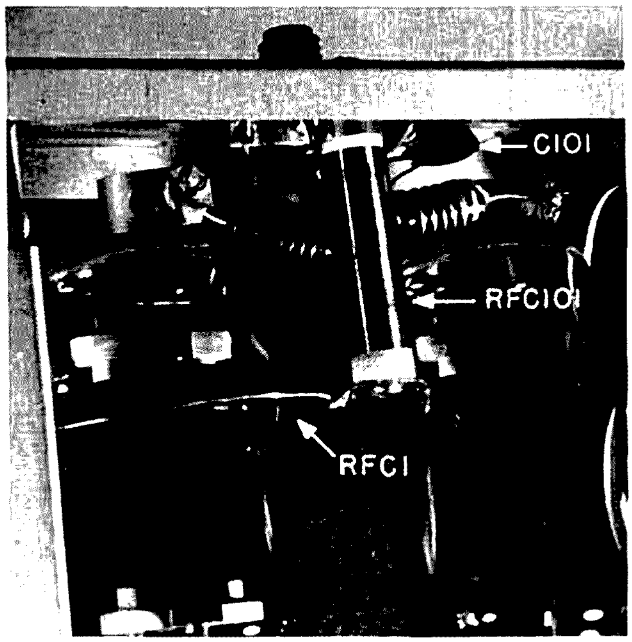

fig. 2. High-voltage compartment of the HA-14 showing the added plate choke and bypass capacitor.

This input assembly, which is labeled L101 in fig. 1, is resonated to 160 meters by capacitor C102 (three 300-pF and one 200-pF mica capacitors in parallel). Do not try to get away with too few capacitors here as there is an appreciable rf current flow. The third winding on the bifilar choke consists of 6½ turns of number-20 (0.8mm) stranded, insulated wire, held in place by Duco cement.

measuring dc power input

The HA-14 has no provision for measuring either plate voltage or plate current, but this is easily corrected. Whatever the merits the vswr bridge may have had when the HA-14 was used

in an automobile, for fixed-base operations everyone should have a decent vswr bridge. The HA-14 meter has six divisions which readily permits full-scale readings of 6000 volts and 600 mA. The vswr function was abandoned.

The vswr switch and variable resistor were removed, a new rotary two-pole two-position deck switch was installed, and a meter circuit arranged for monitoring dc power input. Since another wire is to be added to the power cable to measure plate current with the amplifier meter, a new cable is needed. I used a cable four-feet (1.2m) long with number-14 (1.6mm) wires for the ac and filament lines.

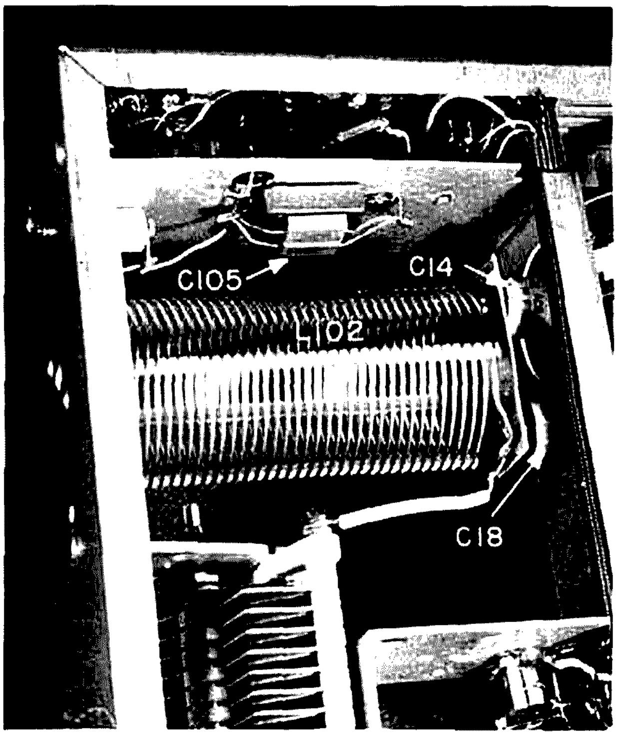

fig. 3. Pi-network compartment of the 160-meter Heathkit HA-14.

Plate current is measured by inserting a one-ohm resistor in the high-voltage ground return in the Heath HP-24 power supply and using the amplifier panel meter as a voltmeter to measure the voltage drop across the resistor. Open the ground return lead of diode CR14 and insert the one-ohm resistor. Since I didn't have a one-ohm resistor, I used ten 10-ohm resistors in parallel. Fig. 5 shows the location of the resistor bank, R201, in the circuit. The hot end of the resistor is then wired to pin 1 of the HP-24 octal power socket and then through the new power cable to pin 4 of the HA-14 connector. Within the HA-14 a wire is run from pin 4 on the power socket through the cable bundle to R101 and then to S101.

The HA-14 ALC threshold level is set by a resistive voltage divider across the high-voltage line (R1 thru R7) in the HP-24 power supply. A level of approximately 6 volts dc is brought through pin 9 to the right-hand terminal of the strip seen in fig. 4 and makes an ideal point to use for measuring plate voltage. From that point a 10k resistor, R102, goes to switch S101; S101 is used to switch between plate voltage and plate current readings. While a dpdt toggle switch could be used for S101, I used a rotary switch so the front panel of the HA-14 would retain its original appearance.

Blinking of the pilot light during modulation or keying is cured by the addition of capacitor C104. More effective rf bypassing at the bifilar choke was accomplished by adding C103. To retain the panel's original appearance, an old variable resistor from my junk box was used to fill the hole left by the removal of the bandswitch.

test procedure

The initial testing of an rf amplifier of this power level requires a good 50-ohm dummy load capable of handling the power. Vswr bridges at both the input and output are essential. It is highly recommended that an rf power output meter be used if overall performance and efficiency are to be determined.

With minimum power output from the exciter, turn on the HA-14. There should be no mistaking the closure of the HA-14 antenna relay when you key the exciter. Rapidly turn the tank capacitor, C15, to resonance as indicated

by maximum power output. C15 should be about two-thirds meshed at the 1800-kHz end of the band and near minimum at the 2000-kHz end.

Next test the dc power metering. Because of the high voltage involved, these tests can be extremely dangerous and should never be made without a responsible and knowledgeable person in the shack with you. From Heath specifications and my own tests, the no-load voltage on the power supply is about 2500 volts dc. Slight changes in the value of R102 may be necessary to bring the panel meter to a 2500 volt reading (2.5 on the swr scale).

With a reasonably accurate 500 mA meter in series with the high-voltage cable, increase the drive from the exciter to produce 200 mA current flow in the HA-14. The HA-14 panel should indicate 200 mA (2 on the swr scale). A slight adjustment in the value of R101 may be necessary to bring the HA-14 meter into agreement.

Next test the input circuit. A vswr bridge between the exciter and the HA-14 should indicate a vswr of 1:1 when the exciter is keyed and the HA-14 is putting out power. If this is not the case, temporarily solder another 100 pF capacitor across C102. If the vswr improves, it means that C102 is not large enough; if the vswr worsens, it means C102 is too large and the 200 pF should be reduced 100 pF. If you find you cannot bring the input vswr to 1:1 by

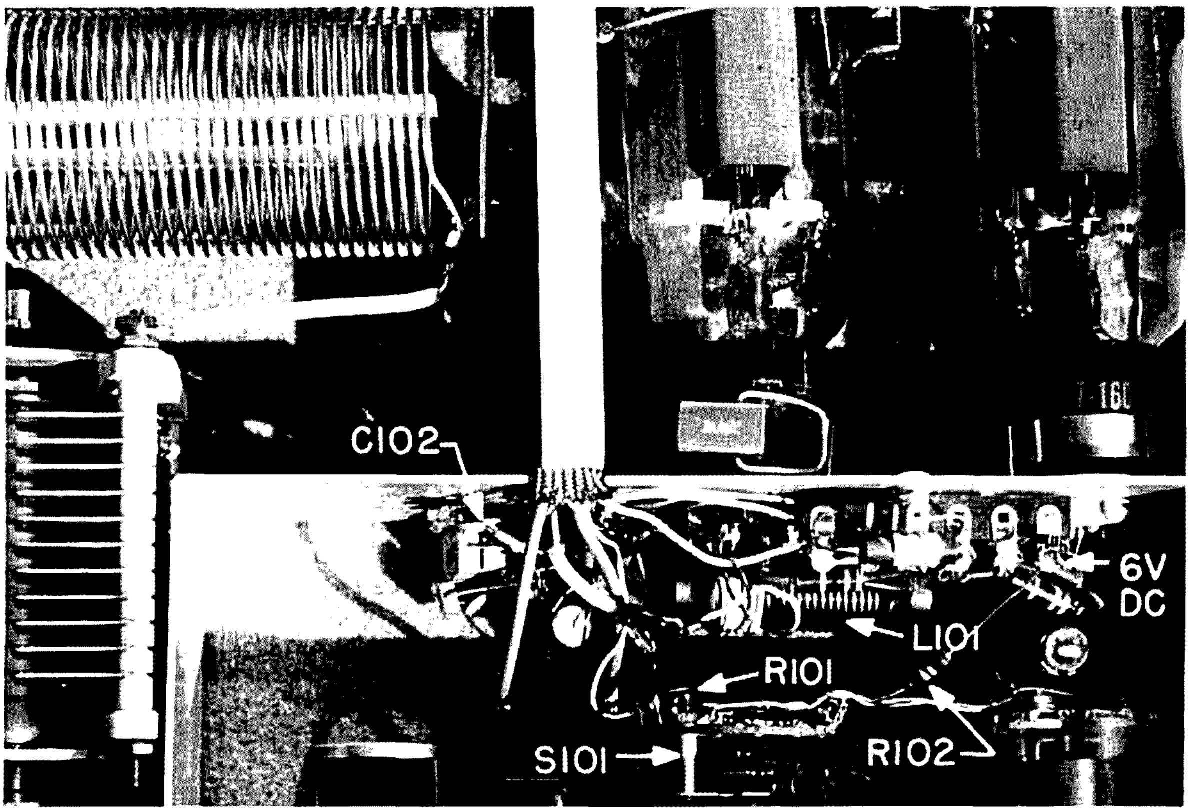

fig. 4. Input network and metering circuit.

adjusting the value of C102, use that value at C102 which brings the vswr to a minimum and work on the third winding of L101, adjusting the number of turns until the vswr is 1:1.

When the input vswr is satisfactory and the output circuit is tuned to resonance, increase exciter drive until the HA-14 plate current is 250 mA; this is 500 watts input. Output power is about 260 watts for an efficiency of about 52 per cent. Amplifier power gain is about 6.7 dB (55 watts drive for 260 watts

output). Increasing exciter drive until the HA-14 plate current is 400 mA produces approximately 800 watts input (the key should be held down no more than 5 seconds). Power output is 450 watts for an efficiency of about 56 per cent. Power gain is about 7 dB (90 watts drive required for 450 watts output).

Now you can connect your antenna. However, the antenna should have reasonably good vswr. If your antenna is not reasonably flat, you are going to run into problems. The variable tank capacitor, C15, because of its limited range, may not be able to bring the tank circuit into resonance if the amplifier load is too far off from 50 ohms resistive.

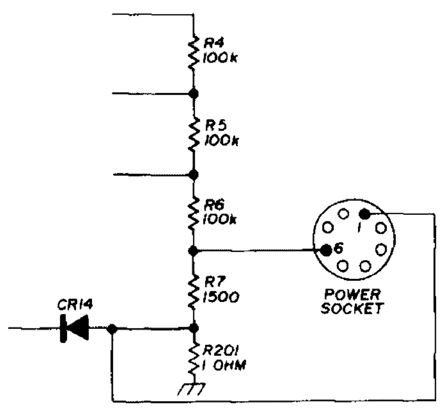

fig. 5. Partial schematic of the Heath HP-24 power supply showing the addition of resistor R201 to permit plate-current monitoring.

Heath SB-200 modifications

The rf circuits in the HA-14 and SB-200 are almost identical. While the HA-14 has C18 (a 350 pF, 5 kV capacitor), the SB-200 has no such capacitor. It can be obtained from a regular parts supplier. Heath is the source of the choke RFC101 (Heath part 45-61, 50 μH rf choke) used in both the HA-14 and the SB-200 160-meter modifications.

Because of the extensive metering in the SB-200, the metering changes described for the HA-14 are not required. This saves time and parts when modifying the SB-200. In addition, the vswr bridge is retained.

A variable pi-network output capacitor is used in the SB-200. While a fixed 2000 pF mica capacitor must be added to the output network as in the HA-14, the variable capacitor in the SB-200 should give some added control over antenna loading.

I have not modified a SB-200 but an inspection indicates the following items should be considered: C11 and C19 (bifilar choke bypasses) should be increased to 0.05 μF. Capacitor C12 may have to be removed. While C102 in the HA-14 has one side grounded, in the SB-200 this 1100 pF capacitor should be placed directly across one of the windings of the bifilar choke.

The bench checks of amplifier gain were substantiated on the air. There were not many reports as daytime activity on 160 meters is quite low. Nevertheless, during one major contest an ssb contact with Hawaii occurred at sunrise with the amplifier in use – my signal was not readable while I was running barefoot.

I estimate that no more than six hours are required to make the modifications to the HA-14 and less for the SB-200. Except for the plate choke RFC101 (which is available from Heath) all parts are quite standard. Parts cost is less than $15 if all parts are purchased new, the most expensive item being the tank coil, L102. For the SB-200 modification only the 350 pF, 5 kV, capacitor needs to be obtained. No attempt should be made to run the HA-14 at 1000 watts input on CW – the tubes would probably collapse.

references

1. Douglas Stivison, WA1KWJ, "Caribbean 160-meter DXpedition," *Worldradio News*, August, 1972, page 8.
2. Irvin M. Hoff, W6FFC, "High-Frequency Power Amplifier Pi-Network Design," *ham radio*, September, 1972, page 6.

ham radio

vhf fm receiver alignment techniques

Complete discussion of alignment techniques for the three most popular vhf fm receiver circuits

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There are two seemingly complementary diseases which afflict large numbers of amateurs: "alignaphobia" and "alignatosis". The first of these is characterized by an absolute terror of making any attempt to align any electronic circuit. The etiology of the disease probably lies in a very conservative (e.g. *don't touch*!) early education in electronics reinforced by a lack of self confidence, the kind nurtured best by some mystical commodity usually called "experience." The latter disease, on the other hand, causes the victim to inexplicably twist, turn and adjust everything in sight with neither rhyme, reason nor clearly defined purpose. Which is to be most feared is best left to the philosophers. The purpose of this article is to offer a little insight which will help both sufferers along; at least where fm receivers are concerned.

typical fm receivers

Knowledge has a way of alleviating both forms of alignment syndrome provided it is supplied in big enough doses. This need not be frightening as the proper dosage is surprisingly close to that level required to pass the technician/general class examination. Toward that end let us consider first a couple of the more popular fm superheterodyne designs.

Fig. 1 shows the block diagram of one typical superheterodyne fm receiver. The front end converts the 146-MHz vhf fm signals to a lower, more manageable intermediate frequency through a process of heterodyning the rf signal against that from a local oscillator. Most amateur receivers or

transceivers use three to five transistorized i-f amplifier stages and, sometimes, a limiter. Another popular alternative, providing at least as much gain, is the use of one or two very high-gain IC *gain blocks* of which there are several readily available.

The final stage, prior to the audio section, is the fm demodulator. This stage processes the frequency-modulated i-f signal to extract the original audio information. In some receivers it also serves to supply the dc feedback control voltage used to drive any automatic frequency control (afc) circuits which might be used in the receiver.

Many fm receivers are actually dual-conversion jobs such as shown in the block diagram of **fig. 2**. In this system the front end converts the rf signal to the same high i-f (usually 10.7 MHz although other frequencies are sometimes used) as used in single-conversion designs. This signal is then heterodyned against the output of a crystal oscillator to produce a low i-f in the vlf range (typically 455 kHz).

The problem of aligning an fm receiver is primarily in learning to recognize detector types and knowing the proper procedure for that type with which you are confronted. There are three basic types of fm detectors in general use: discriminator (Foster-Seeley), ratio detector, and quadrature detector. I recognize that others exist but not in sufficient incidence to warrant coverage here.

Fig. 3 shows a simplified but essentially complete version of the Foster-Seeley fm discriminator. In this circuit signal voltages from the primary of T1 are added to signal voltages in the secondary. When the signal frequency is exactly equal to the frequency to which the secondary is tuned the output voltage will be zero. Deviation, whether caused by a shift in carrier frequency or by the process of frequency modula-

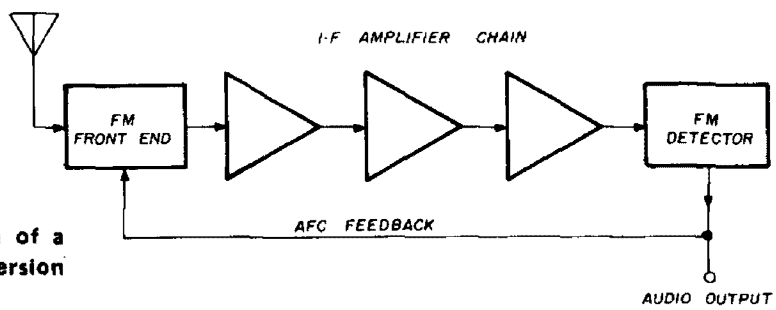


fig. 1. Block diagram of a typical single-conversion fm receiver.

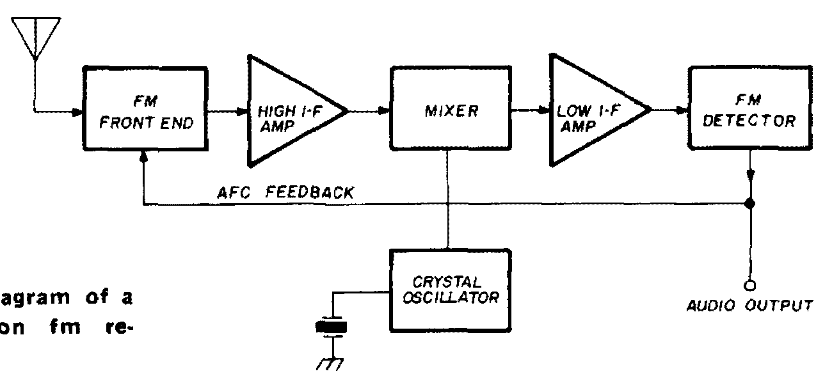


fig. 2. Block diagram of a double-conversion fm receiver.

tion, will produce a positive or negative output voltage depending upon the direction of frequency shift. This varying output voltage is the audio signal.

Another traditional fm detector is the ratio circuit of **fig. 4**. There are two immediate characteristics of this circuit which distinguish it from the Foster-Seeley discriminator: the diodes point in opposite directions and the circuit *includes an electrolytic capacitor* (C3). The capacitor is sometimes referred to as an a-m suppression capacitor as it bypasses amplitude variations sufficiently to reduce the need for a limiter. In the ratio circuit the relative charges on capacitors C1 and C2 will have a 2:1 ratio when the i-f frequency is precisely equal to the resonant frequency of the secondary of the transformer. Frequency deviation causes the ratio to change, resulting in an audio output signal.

quadrature detectors

The last type of detector to be considered here is an old friend which has been given a nominal re-birth by the advent of integrated-circuit technology: the quadrature detector. This circuit gets its name from the fact that it

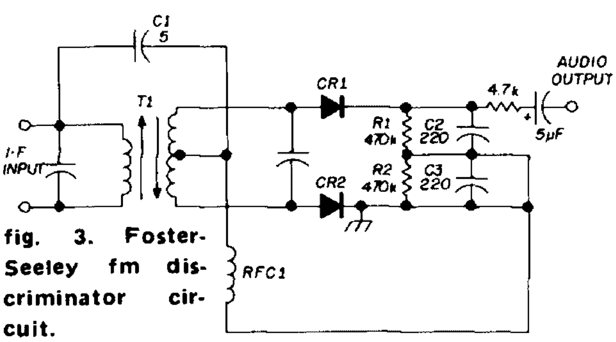


fig. 3. Foster-Seeley fm discriminator circuit.

demodulates an fm signal by combining two versions of the i-f which have a phase difference of 90° (that is, the signals are "in quadrature"). An example of the IC Quadrature Detector (ICQD) is shown in block form in **fig. 5** and as a schematic in **fig. 6**.

The input stages to the quadrature detector circuit are wideband, high-gain amplifiers which serve to limit the amplitude variations of the input signal. This eliminates a-m components such as

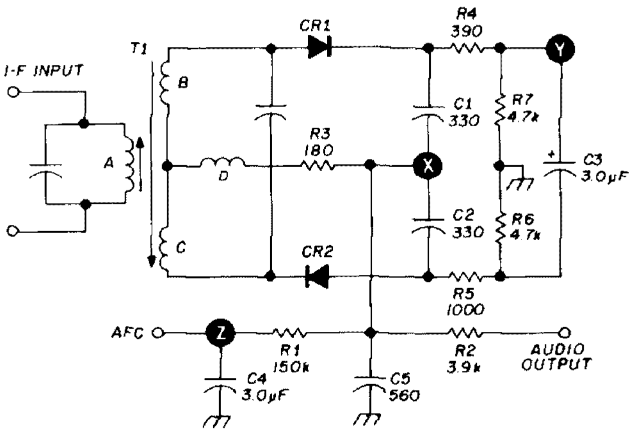


fig. 4. Fm ratio detector is found in many fm receivers.

noise and converts the signal to a series of squarewaves. These square waves have varying periods and durations, the actual value of which depends upon the input frequency and the nature of the modulation.

The square waves from the limiting amplifiers are fed to a splitter section which separates them into two channels. One is fed directly to the synchronous-gated detector while the other is fed to an external (see **fig. 6**) 90° phase shift network. The shifted version is then fed to the alternate input of the gated detector.

One process of the gated detector is to integrate the detector output pulses *to extract the audio signal*. Be aware that the use of an IC in the detector does not automatically indicate the use of a quadrature circuit. The real telling feature is the phase coil in place of the transformer. There are many types of ICs used in fm i-f stages which also include the detector diodes. An example of such a device is the popular RCA CA3043. These devices are given away by the use of a transformer.

alignment instruments

Fm alignment procedures all require some sort of controlled signal source as the standard. Some amateurs may have a complete fm alignment laboratory that would make the GE and Motorola two-way radio R&D labs jealous, but such resources are not needed, as will be demonstrated. However, if you have the resources to select a reasonably high grade instrument there are several factors to look for:

1. Low residual signal leakage. It is not very helpful if the signal level leaking around the attenuator or through the cabinet flanges can drive your receiver into hard limiting.

2. Reasonably accurate frequency dial or a means of setting to accurate points (i.e. crystal calibrator or a high-level output for a frequency counter).

3. Calibrated output level. The usual method is to have a meter with a known *set point* preceding the attenuator. The attenuator is then calibrated in microvolts or dBm.

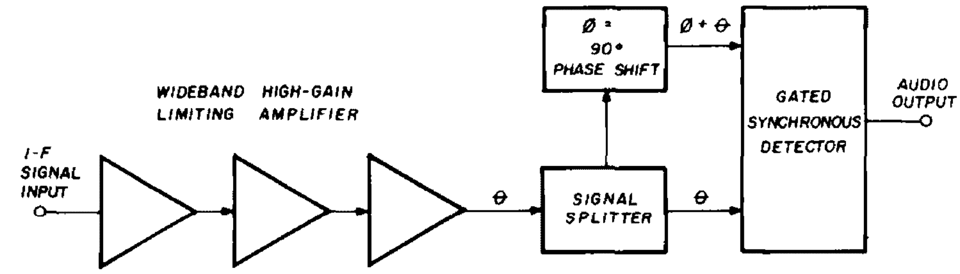


fig. 5. Block diagram of the MC1357P IC fm quadrature detector.

4. Control over modulation level and some means of indicating amounts. If an a-m only signal generator is being considered, a means is needed for turning off the modulation.

5. Reasonable short-term stability and rugged construction.

There are a number of signal generators available which will suit the needs of the amateur. Some are available at attractive prices on the surplus market and include the venerable Measurements series (models 60, 80, etc), the surplus TS-497 (military version of the Measurements 80), and the old Boonton 202. Fig. 7 shows an updated version of the model 202 now offered by the Boonton

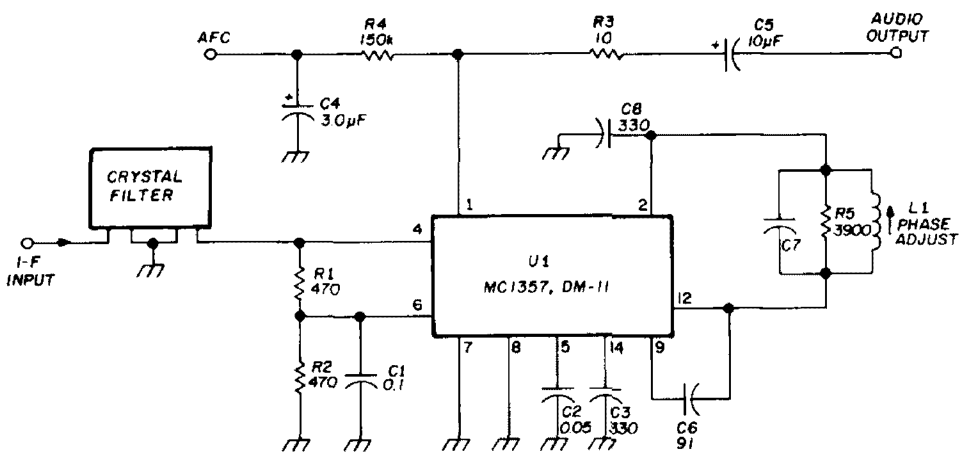


fig. 6. Typical circuitry used with the fm IC quadrature detector.

division of Hewlett-Packard as their model 202H. The main signal generator produces a high quality vhf fm signal at a calibrated output *level*.

The *Univerter* in fig. 8 takes the output of the 202H and heterodynes it down to lower frequencies for use in i-f alignment. One feature of the Univerter is that it produces a low signal with the same deviation (because it heterodynes rather than divides) and output level as the 202H. This allows the calibrated output to be controlled by the 202H attenuator.

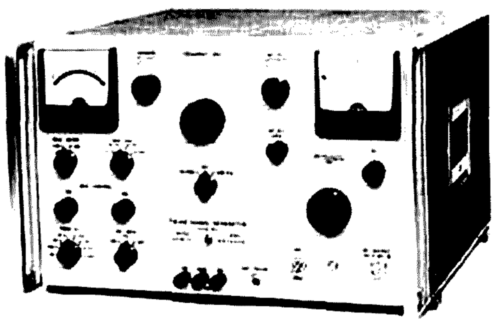


fig. 7. Hewlett-Packard model 202H fm signal generator (photo courtesy Hewlett-Packard).

Fig. 9, the Measurements model 800, is an updated fm signal generator. A block diagram of the model 800 is shown in fig. 10. This instrument has proven popular with the commercial vhf fm mobile-radio crowd as it is reasonably portable for installation in a crowded service truck.

As amateurs on limited budgets we often find it necessary and advisable to have available certain contingencies which allow a goal to be realized. While I will readily concede that a kilobuck signal generator might be the *best* way to align an fm receiver, I think it is necessary to offer a viable alternative to those whose resources are limited to a vtvm or vom and a junk box full of parts. If you fall into this category, as most of us do, the signal generators shown in fig. 11 and fig. 12 are for you.

Fig. 11 is a simple two-transistor crystal *oscillator which should oscillate* between 1 and 13 MHz or so, depending upon what type of transistor you use. The transistors can be almost any small-signal type offering good gain at the frequency range of interest. One oscillator I built used some vhf pnp Germanium types salvaged from an old Delco car radio. The crystal, Y1, should be chosen to produce either the desired frequency (in the case of the i-f) or a sub-harmonic of the front-end rf frequency. For example, a 10.7-MHz crystal for the i-f and either 6- or 12-MHz crystals which are sub-harmonics of the receiver frequency. The trimmer C_T can be used to zero the crystal frequency. This can be done with a crystal calibrator and a receiver, external counter or another fm receiver known to be correctly tuned to local repeater frequency.

The low cost of TTL IC logic devices means that the calibrator of fig. 12 can be built for practically peanuts. If the crystal is a 500-kHz type the output can be used as 5-kHz markers in a sweep alignment procedure. If you eliminate the two SN7490P decade dividers and plug in a 455-kHz crystal, the circuit can be used to supply the low i-f frequency used by some receivers. This circuit will oscillate, when selected TTL devices are used, up to frequencies of several MHz.

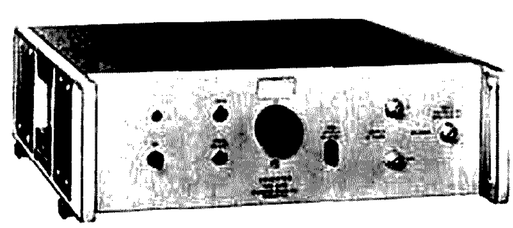


fig. 8. Hewlett-Packard 207H Univerter i-f converter for the model 202H signal generator (photo courtesy Hewlett-Packard).

Please note that the procedures described here, while being reasonably universal, are fairly generalized. They are not an end-all for all receivers. Some radio manufacturers might toss in a few wrinkles of their own just for fun. If you have a service manual or other source which promotes a certain technique as best, bow to their wisdom and follow it if possible. In the absence of manuals, or where resources do not permit, following the procedure prescribed here should prove successful.

A typical sweep alignment set-up is shown in fig. 13. The signal generator provides a calibrated, controlled source which is supposed to effectively simulate the input from the antenna. In some manuals a dummy antenna will be specified for interconnection between the generator and the receiver. The marker is a crystal oscillator which is used to provide pips on the oscilloscope trace to aid in identifying specific frequency points. The adder is an isolator which allows interconnection of all the instruments without undue interactions which might tend to make the job impossible. When the audio output from the receiver is added to the detected output from the other sources the oscilloscope will display what is essentially a calibrated frequency response curve for that receiver.

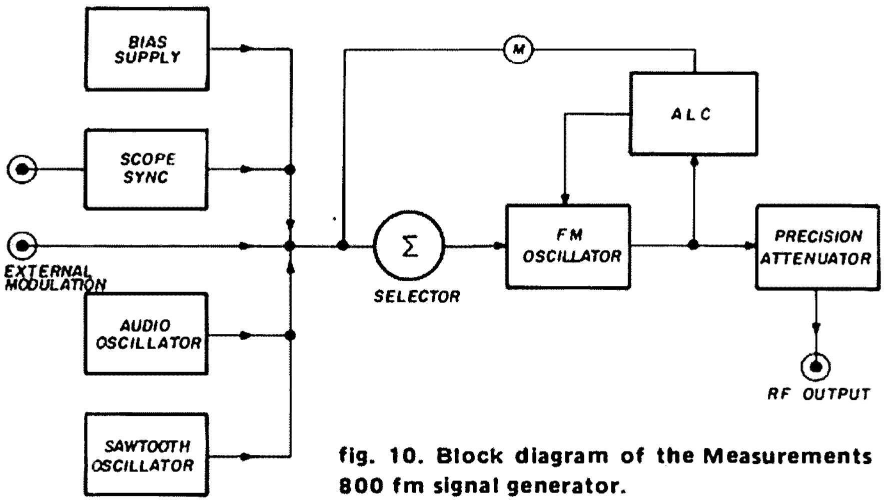


fig. 10. Block diagram of the Measurements 800 fm signal generator.

fig. 9. Measurements model 800A fm signal generator (photo courtesy Measurements Division, Edison Electronics).

Fig. 14 and 15 show the various curves associated with sweep alignment. The trace in fig. 14A is an i-f response curve. The dip at the top end of the curve will be especially noticeable in wideband equipment. Authorities usually claim that the dip should have a depth of ten percent of the overall amplitude. Let that be a maximum. Do not try what one serviceman acquaint-

ance of mine did, and create the dip from what was essentially an almost ideal flat response!

Fig. 14B shows the discriminator output curve. Note that the output voltage goes through zero when the input signal frequency is equal to the resonant frequency of the discriminator transformer secondary. This should give you some ideas as to how an fm deviation meter operates. The curve in **fig. 15** is a fairly typical response curve showing a marker pip at the center frequency.

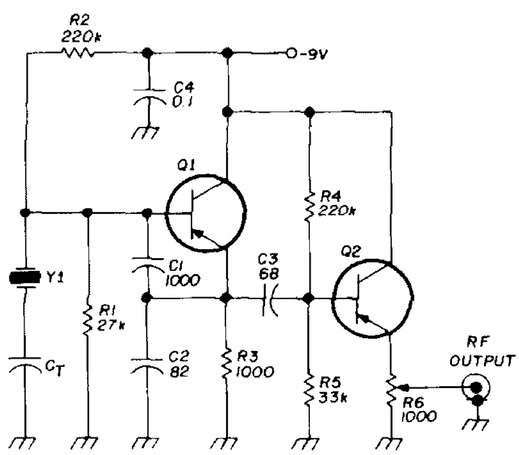


fig. 11. Simple two-transistor alignment oscillator.

Most receiver manufacturers clearly state in their literature exactly where they want the i-f alignment signal injected. This might be a jack or test point in the tuner or in the input section of the i-f amplifier strip. Follow their advice if the point is known. In the absence of good data connect the signal generator output to either a capacitor connected to the mixer input or a "gimmick" dropped inside the first i-f transformer. The gimmick, in this context, is simply a short length of *insulated* hook-up wire. About ¼-inch (6-mm) is bared on one end to make contact with the generator output cable.

One common procedure calls for a zero-center voltmeter (most vtvms and fet voltmeters can be made zero center by adjusting the zero control) to a point such as Z in **fig. 4**. Apply an fm signal to the input and adjust the secondary of the detector transformer to zero volt. The meter will shift positive on one side of the correct setting and negative on the other side. Now adjust the i-f tuning to produce a curve such as that shown in **fig. 15**. During this operation keep the input signal level well below the limiting point but above the noise to eliminate any ambiguity. Overdriving the receiver (driving it into limiting) has a tendency to broaden the response and obscure the true peak.

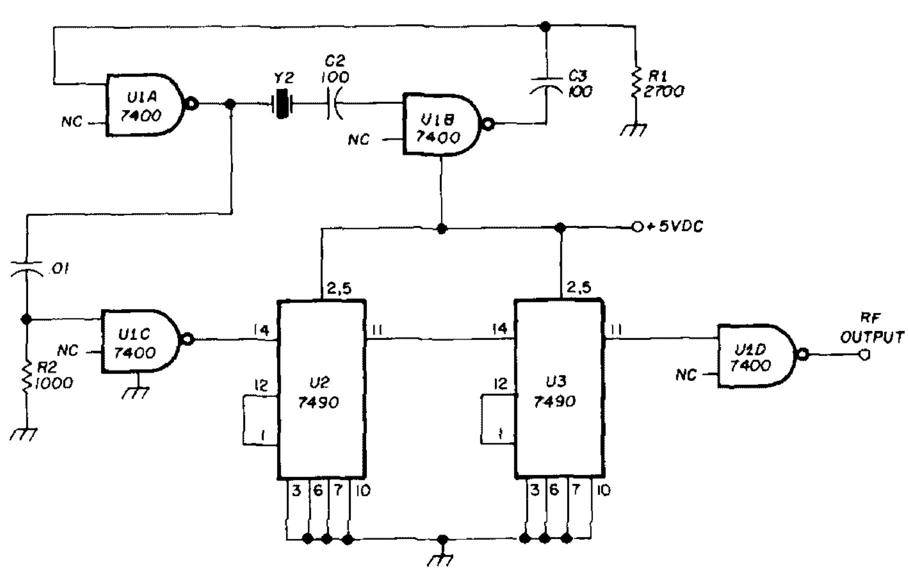


fig. 12. Simple TTL digital logic IC alignment oscillator.

One aspect of the procedure for aligning the secondary of the transformer is that the correct point is established with the least total harmonic distortion. Although that is not the best way to go (unless the proper THD analyzer is available) I have known professional servicemen who could rough in an alignment off the air to an extent that the operator would not know the difference.

The local oscillator in the front end can be set by adjusting the trimmer capacitor across each channel crystal for a zero at point Z when a precisely known input signal is applied to the input. The signal can be from a multi-kilodollar synthesizer/frequency meter, a close-at-hand transmitter, simple crystal oscillator as shown earlier (provided it is accurate) or from your local repeater. In my Heath HW-202 the transmitter oscillator is used for this purpose. Again, this assumes that some other means was used to verify the transmitter's correctness.

Peak the rf amplifier tuning to either a channel specified by the radio manufacturer or a channel approximately mid-way in the group of channels used. Of course, if the rig is a single-channel affair peak it to that channel. Rf amplifiers will typically have two adjustments. For both it is necessary to keep the input signal level down to a barely audible level. If the input is a common base or gate design one adjustment will be noticeably broader than the other.

non-swept alignment

The use of the transmitter oscillator applied to the receiver alignment in at least one model transceiver implies that it is not absolutely necessary to use a sweep generator. In fact, the non-swept technique, properly applied, will yield comparable results. However, it must be noted that sweep alignment is preferred by high-fidelity buffs because of the interaction of the i-f alignment and the channel bandwidth. In lower grade fm broadcast receivers and narrow-band equipment the need is less acute.

It is important to note, however, that the signal source must be unmodulated. If your source is an a-m signal generator, turn the modulation off. This can be done by turning the modulation control switch to *off* or *external*. Do not depend upon the modulation level or percentage control to sufficiently reduce the level.

As was true in the swept technique, adjust the detector transformer secondary to null when the unmodulated signal, at the i-f of a channel frequency, is present at the input. Coupling to the i-f strip is the same as for the swept technique. Connect a high-impedance dc voltmeter to a point in the receiver

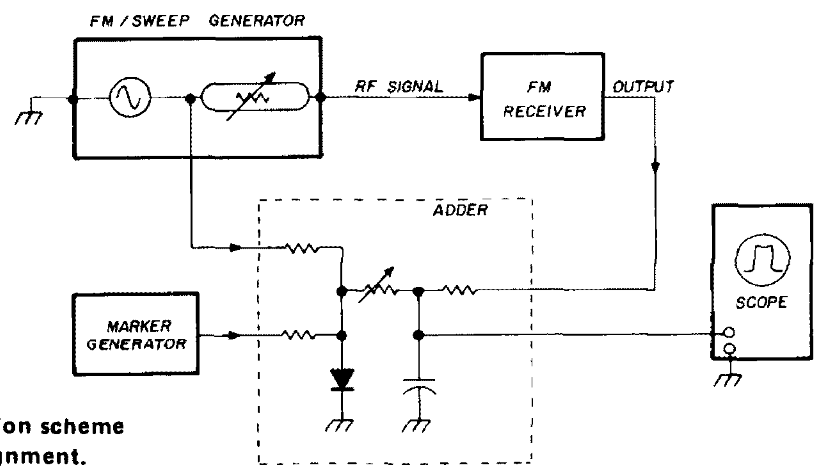


fig. 13. Connection scheme for fm sweep alignment.

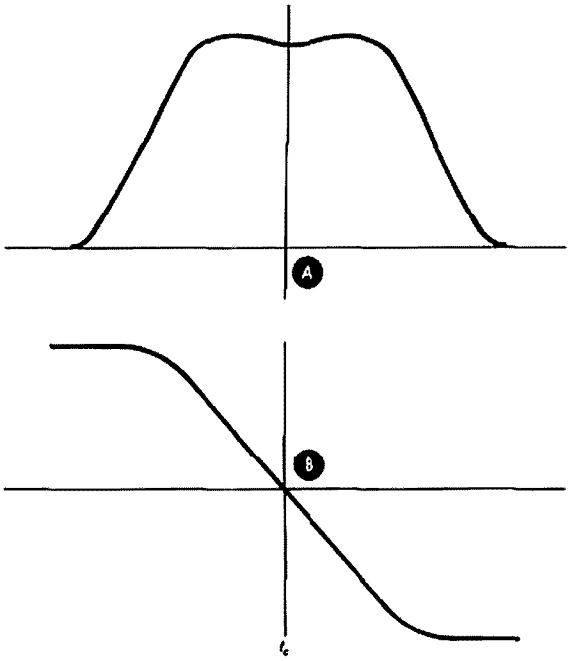


fig. 14. Typical alignment curves from sweep alignment of fm receiver.

which produces a voltage proportional to the strength of the input signal. Point Y in fig. 4 is such a point. In vacuum-tube sets the grid of the limiter does nicely while in transistorized types the emitter resistor of the limiter is good.

Another method uses an rf detector probe at the limiter input although care must be exercised not to detune the output *i*-f transformer. It is necessary that the input signal level be kept low. In some cases it is a good idea to disable the agc system to prevent interaction with the alignment. No good advice can be given on this point. In some receivers you merely ground the agc line. In others you must apply a fixed bias of one polarity or the other, while in still others it is best to physically interrupt the agc line (some receivers have removable jumpers specifically for this purpose). *Do not* attempt to use your ears as a monitor of the quieting level as it will probably be inaccurate.

In both the swept and non-swept alignment procedures start at the detector and work toward the front-end. Go through the procedure several times looking for a slight improvement each time. This optimization is needed because there will be a slight interaction between adjustments.

One admonition: *Don't overtighten i-f slugs and trimmer screws.* Many ferrite slugs will become seized if they are tightened against either the top or bottom stops. Additional force will tend to make them break. One sure sign that this has occurred is a "crumbly" feel as you adjust. Stuck slugs can often be freed by application of a slight amount of heat.

I-f transformers which do not tune usually indicate one of two conditions: the transformer is defective and requires repair or replacement, or that coil is actually in the transmitter circuitry (identify *all* parts before making any adjustments). Note that many a good rig has had to make a trip to the repairman because of misguided alignment attempts. Believe me, the service guy can almost always tell a set that has been tinkered with.

quadrature detector alignment

Aligning circuits using the MC1357P or Delco DM-11 ICQD detectors requires a special technique. If a swept generator is available the job is simpler. Connect an ac vtvm across the output (in the absence of an ac vtvm a rectified dc type or an oscilloscope will be adequate) and adjust the 90° phase coil

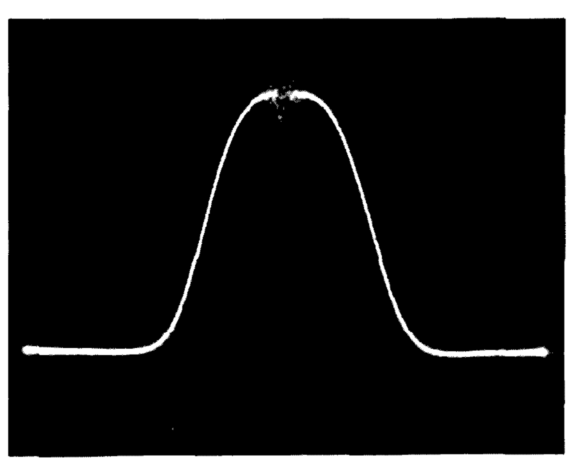


fig. 15. Oscilloscope trace of i-f passband curve with frequency marker.

for maximum output signal level. Connect a dc vtvm through an rf detector probe such as the RCA WG-301 to pin 10 of the MC1357 or DM-11 detector (it is worth noting that if you need a MC1357P replacement and can't locate one, drop into the nearest authorized Delco car radio shop and buy a DM-11 or DM-31). Now you can peak the i-f coils for maximum.

When using an unmodulated signal generator you can approximate the pro-

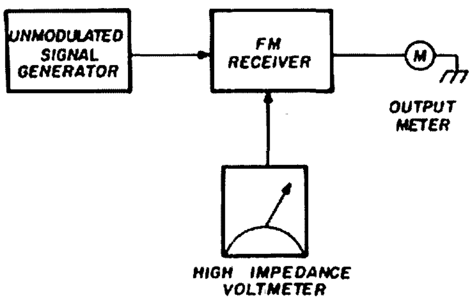


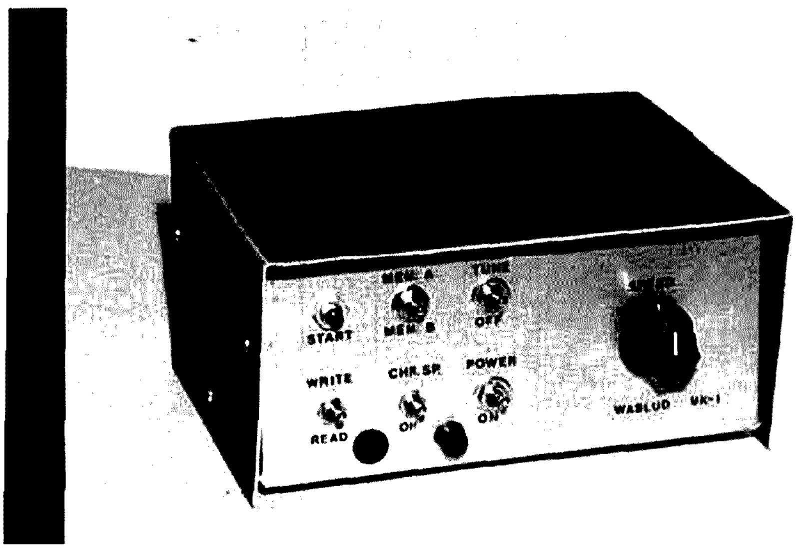
fig. 16. Test equipment arrangement for non-swept alignment of fm receivers.

per phase-coil adjustment by nulling the noise level. The phase coil is correctly adjusted when it is tuned to a null located between two relatively high amplitude noise peaks. However, this is only an approximation as the null tends to be rather broad.

summary

As stated earlier these techniques are rather generalized. They will suffice, however, for most jobs without serious performance loss. One parting word of caution: *Do not align anything as a troubleshooting method*. The first and foremost sign of the novice troubleshooter is the galloping "diddle stick." Remember that alignment doesn't change within short spans of time. If your rig suddenly stops working the fault is *not* alignment. Find the fault first. If it is an alignable component, align *only* the replacement part or the repaired original.

ham radio



Photos by WB9DZI

programmable memory

accessory for electronic keyers

Design of an expandable, programmable random-access memory system for use with electronic keyers

Andrew B. White, WA9LUD, 102 Franklin Street, Urbana, Illinois

During the past few years, several circuits have been published for CW keyers, both with and without programmable memories. Most designers, however, seem to either combine memory capability with a mediocre keyer circuit or design a good keyer with no memory. With the inexpensive TTL ICs and mos memories that are now available, I felt there was room for a better design which incorporated a good iambic keyer with a programmable memory.

Riley[1] had an interesting application of recirculating memories, but the programming seemed awkward, the messages (if more than one) were stored in series, and I felt separate speed controls for the keyer and the memory should be avoided. Gordon[2] used two mos random-access memories to provide two selectable message stores, but the associated keyer was not acceptable. The other drawback I saw in both these approaches was the inability to re-start the read or write sequence until the current message was completed. The design presented here combines an excellent keyer circuit with two programmable memories.

I chose to use the very popular Accu-Keyer circuit described by Garrett[3] as the basic keyer because it features dot

and dash memories, iambic operation, automatic character spacing and low cost. The memory accessory, however, could easily be adapted to many other keyer circuits. For the mos memories, I used the readily available National 1101 (or Signetics 2501) static mos RAM. These memories have a three-state output that enables two or more memory outputs to be connected in parallel with no additional circuitry.

places a zero (TTL positive logic) on the CS input of the desired memory. To read the memory, the *read* line is allowed to float high by switch S3 which enables U7C, gating the data output from the memories to the output keying circuit and sidetone. To write data into a memory, S3 takes the read line to zero, enabling a 500 ns write command pulse triggered by the trailing edge of the clock pulse (see timing diagram, fig. 2).

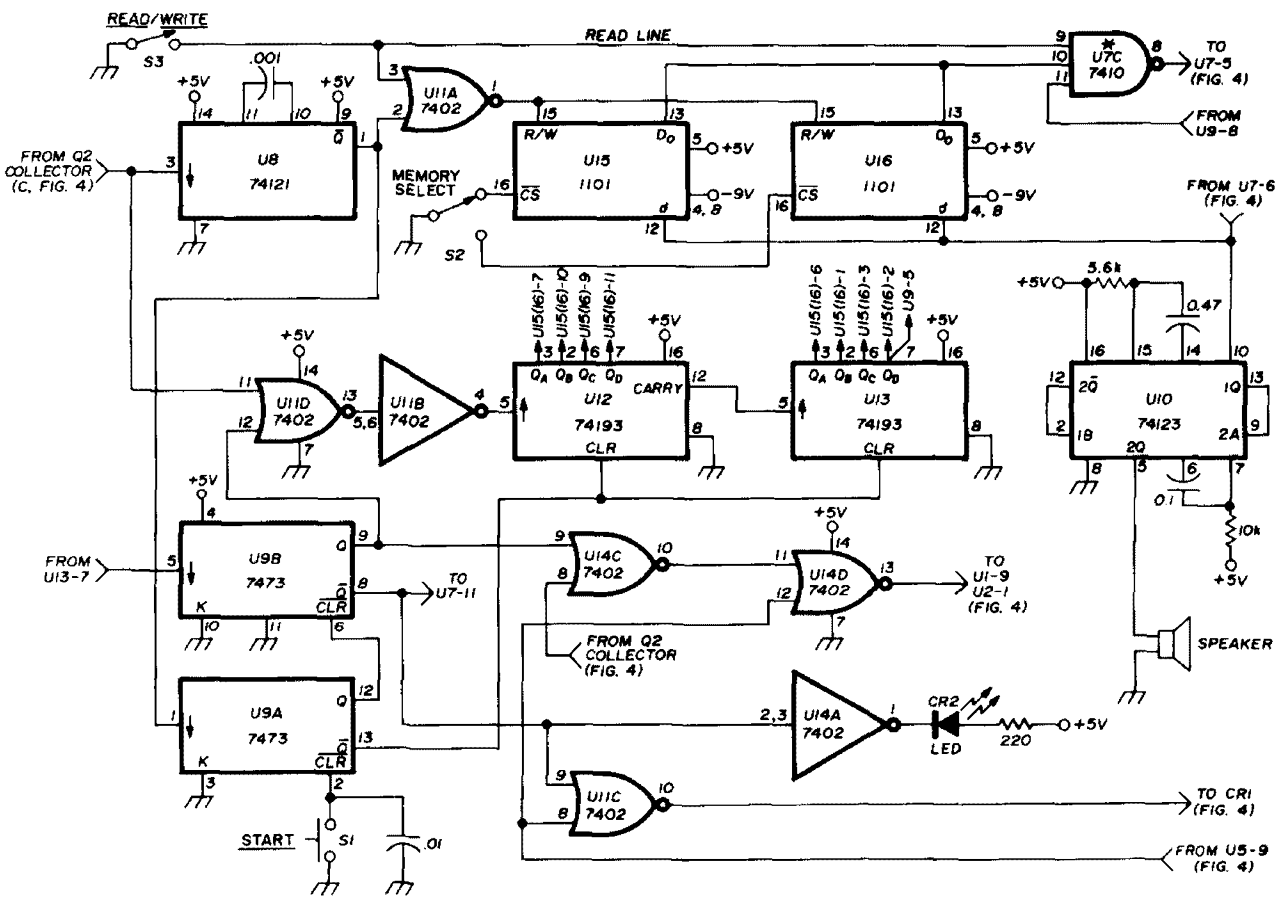


fig. 1. Programmable memory for electronic keyers uses two 256-bit random-access memory ICs, U15 and U16. Timing diagram for this circuit is shown in fig. 2.

memory circuit

Operation of the basic keyer will not be covered here because WB4VVF provided a complete discussion in his original article. Only the operation of the memory circuit itself will be examined. Fig. 1 is a logic diagram of the memory circuit. A memory is selected for either reading or writing by switch S2 which

Both the read and write cycles are initiated by closing S1, a normally-open spst switch, which resets U9A, the start/stop flip-flop U9B, and the counters U12 and U13. When S1 opens, the synchronous four-bit counters count up on the leading edge of the clock pulse from 000 to 255 providing the proper addresses to cycle access to the RAM.

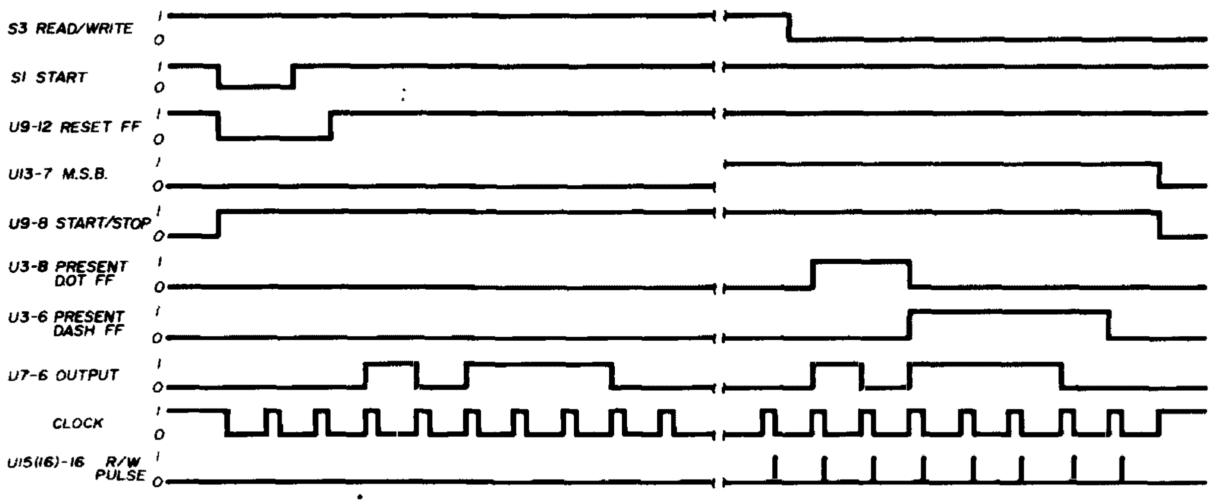


fig. 2. Timing diagram for the programmable CW memory shown in fig. 1.

When the most significant bit of the address goes from 1 to zero, indicating 256 has been attained, U9B toggles, disabling further counting and returning the keyer to normal asynchronous operation.

The start/stop flip-flop, U9B, turns on an LED to signal the user when the keyer is cycling the memory. In addition, a dual monostable, U10, has been included as a sidetone oscillator for programming the memories and use as a code monitor.

Two circuit modifications are necessary to connect the memory accessory to the Accu-Keyer, as shown in fig. 3. The connection between diode CR1, input gate inputs U1, pin 9, and U2, pin 1, and the missing-character flip-flop, U5, pin 8, has to be broken. Since the oscillator must run free during the read and write cycles, the anode of CR1 is connected to U11, pin 10, which is low during the memory sequence. U5, pin 9, (the complement of U5, pin 8) is connected to U11, pin 8, to provide normal operation when not using the memory.

The gate inputs U1, pin 9, and U2, pin 1, are connected to U14, pin 13, so that during the memory write cycle the dot/dash inputs are sampled only when

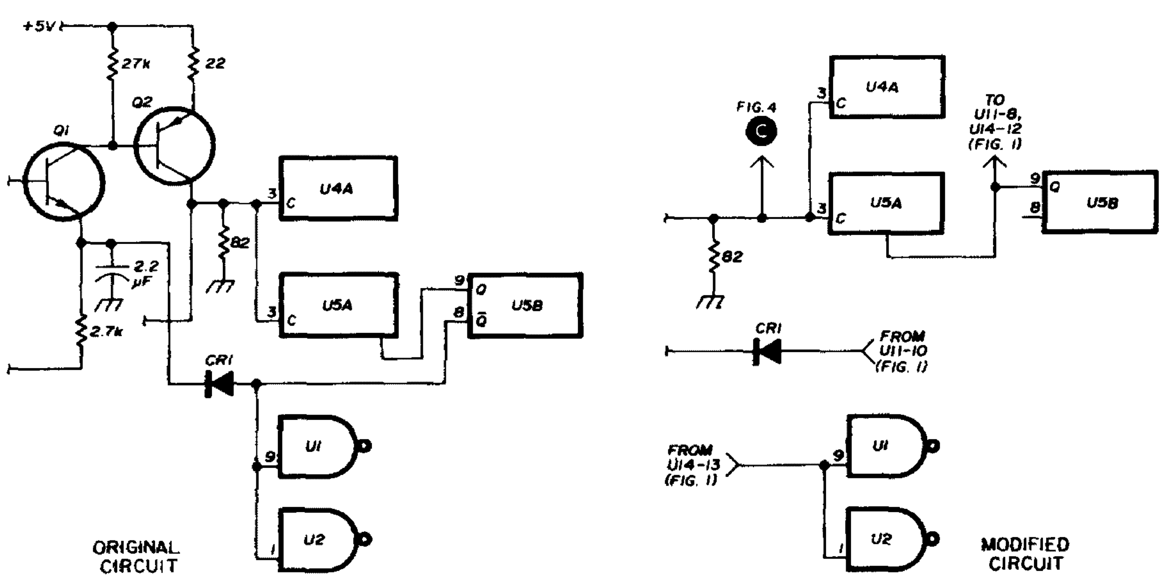


fig. 3. Modifications required to the WB4VVF Accu-Keyer for use with the programmable memory accessory. Complete schematic is shown in fig. 4.

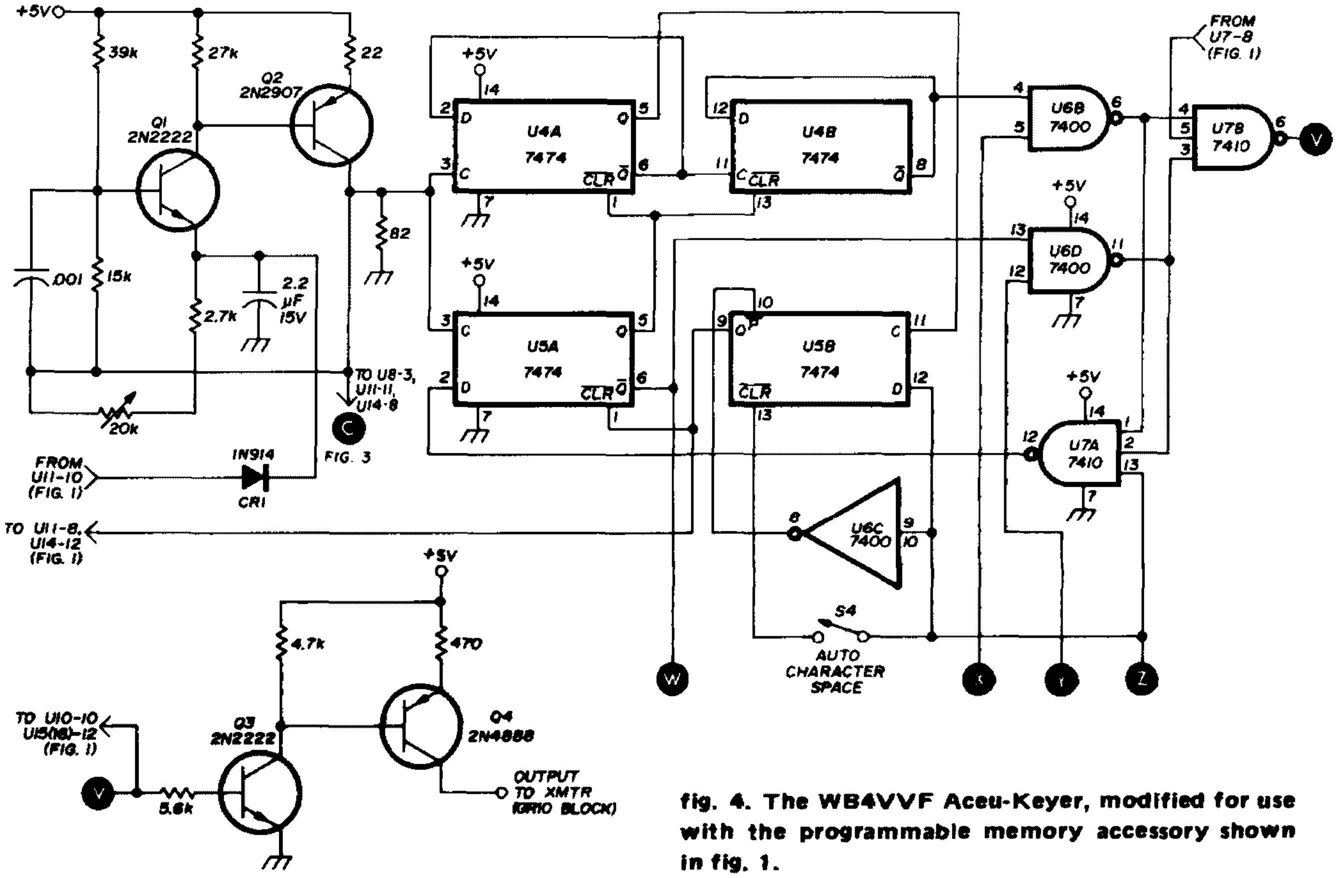
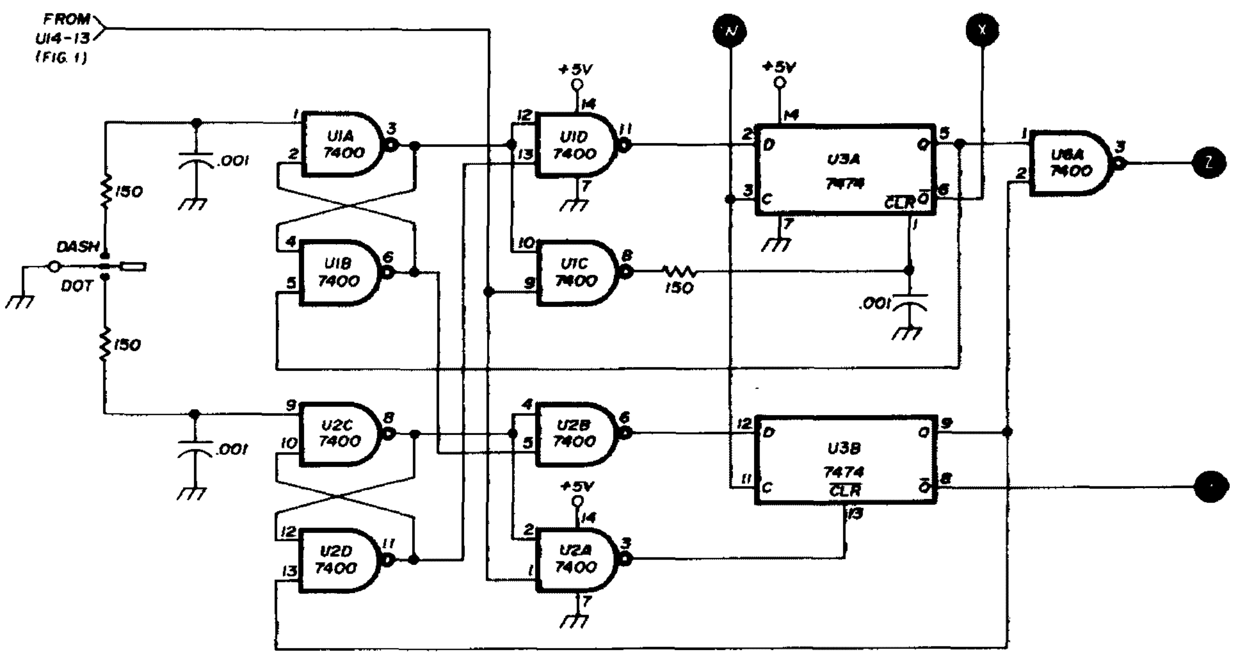


fig. 4. The WB4VVF Accu-Keyer, modified for use with the programmable memory accessory shown in fig. 1.



the clock pulse is high, thus preserving proper dot/dash timing. These inputs are effectively connected to U5, pin 8, otherwise, resulting in normal keyer operation when not using the memory.

The clock in the Accu-Keyer provides the clock pulses for the memory circuit (taken from the collector of Q2). The previously unused section of U7 is wired to inhibit data transfer out of the memory except during the read cycle. Finally, U7, pin 6, is taken to the memory circuit to key the sidetone and input data to the memories.

construction

I built the memory circuit and keyer on two 3-inch (76mm) square pieces of single-clad Vector board using breadboard-type fabrication because wiring errors are much easier to repair. Any other method of construction would work equally as well so long as the usual rf bypass techniques are used. Fig. 5 shows the power supply I used for both the keyer and the memory circuit.

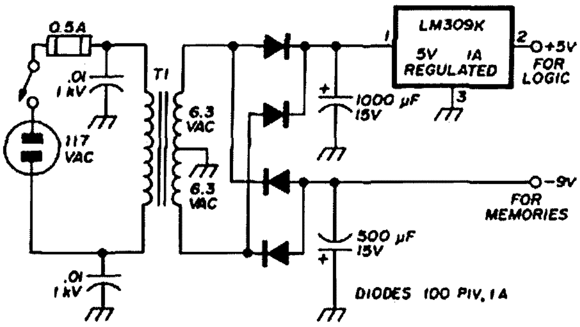
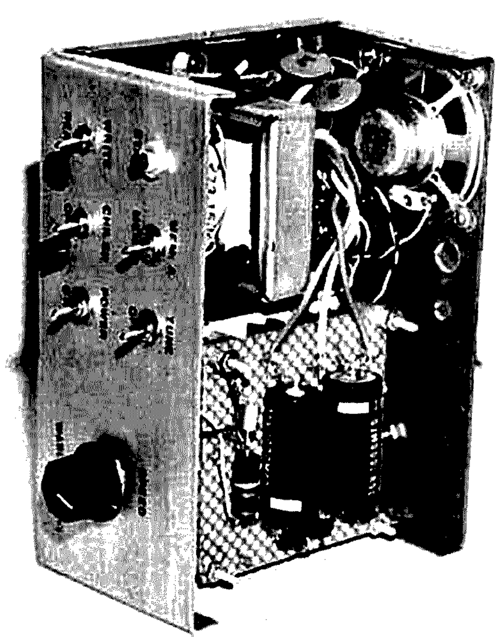


fig. 5. Simple power supply circuit provides regulated +5 volts for the TTL ICs and −9 volts for the random-access memories.

operation

Operation of this memory accessory is quite simple. The operator selects the desired memory with S2. To write data into the memory, S3 is closed, the start button, S1, is pressed, and, when released any data being sent will also be written into the selected memory. If a mistake is made, simply depress the start button *again* and start over. Even if the previous cycle is not complete, the counters will be reset to 000.



Inside view of the memory keyer showing the circuit board stack. Accu-Keyer is on bottom board, memory accessory is on the middle board, and power supply rectifiers and filter capacitors are on top. The +5 volt regulator IC is mounted on heatsink on rear panel.

To read the data in storage, open S3 and press the start button, S1. When S1 is released the memory will cycle through the 256 bits. Two separate messages can be stored and selected by S2. Again, the start button may be repressed at any time to restart the message.

Approximately 30 characters can be stored in a 256-bit RAM. This is enough for the typical Sweepstakes exchange:

B DE WA9LUD/9 64 ILL BK

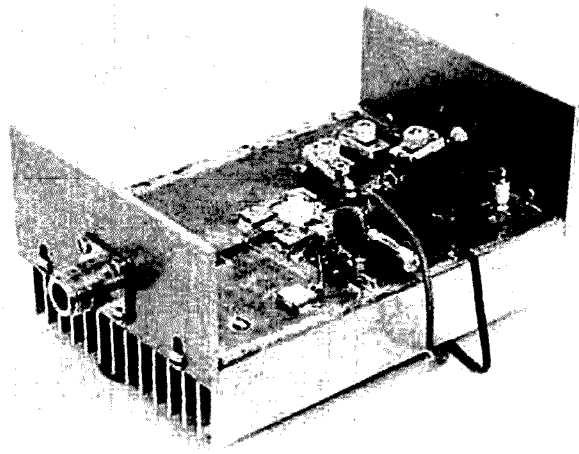
However, the basic circuit can be expanded to a cascade of individual memories for any desired character length.

The entire keyer package cost about $25 to build, but the memory accessory alone can be built for under $10. I have been very pleased with the performance of the unit. The memory replaces a clumsy tape recorder device I used to use during CW contests, resulting in a more relaxed and smooth operation. It has been a *welcome* addition to a crowded operating table.

references

1. Thomas Riley, WA1BYM, "An IC Keyer with Programmable Memory," *QST*, February, 1973, page 26.
2. Michael Gordon, WB9FHC, "Electronic Keyer with Memory," *ham radio*, October, 1973, page 6.
3. James Garrett, WB4VVF, "The WB4VVF Accu-Keyer," *QST*, August, 1973, page 19.

ham radio



solid-state linear power amplifier for 432 MHz

Construction details for a solid-state linear amplifier that provides 10 watts PEP output at 432 MHz

Lance Wilson, WB6QXF, 334 Vista Drive, La Selva, California 95076

Until recently, amateur use of semiconductors on 432 MHz was limited to low-noise preamplifiers, mixers and low-level transmitter stages. Now, with rf power devices developed for the uhf land-mobile service, it is possible to build rf power amplifiers for this band without breaking the bank. The linear amplifier described in this article delivers 10 watts PEP *output* with 10 dB or better power gain, operates from a 12-volt power supply, and is comparable in price to vacuum-tube linears operating at similar power levels. One of the goals in designing this amplifier was to develop a circuit that was rugged with respect to load vswr as well as being easy to build and align — the amplifier presented here meets those goals.

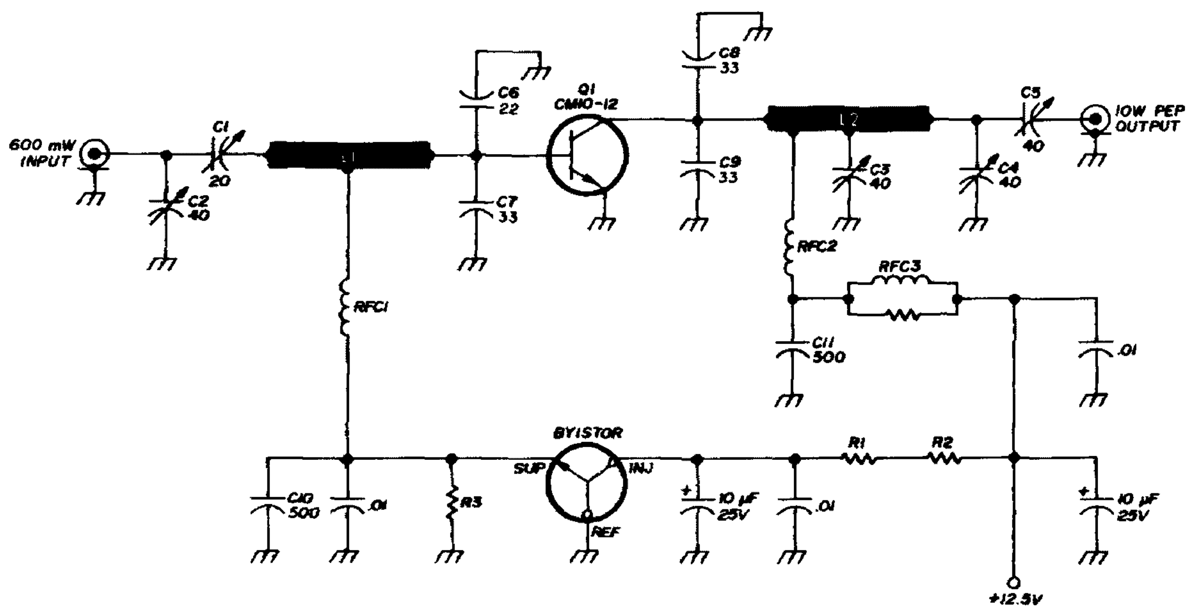
When designing an rf power amplifier, the first task is to select a suitable power transistor. The device I selected is the CTC CM10-12. This device, originally developed for the land-mobile service, is both inexpensive and rugged.* Most solid-state linears for vhf and uhf use 28-volt transistors, but 28-volt devices have two important disadvantages: cost (these parts are always more expensive than 12-volt devices) and power supply

*The CTC CM10-12 is $13.50 in small quantities from Communications Transistor Corporation, 301 Industrial Way, San Carlos, California 94070. The BYI-1 byistor is $7.50. Since the minimum order is $50, this amplifier would make an excellent club project.

availability. Although most amateurs have access to a medium-current 12-volt power supply, 28-volt power presents more of a problem.

Many amateurs find solid-state uhf power amplifiers difficult, if not impossible, to build. Uhf stripline circuits perform exceptionally well, but the Teflon-glass or *duroid* circuit board necessary for their construction is very expensive and difficult to obtain. Furthermore, many amateurs have very little experience with printed-circuit boards. After reviewing this problem, I decided to use lumped constants (mica compression trimmers and conventional inductors). A main consideration in the design of this amplifier was to make it reproducible.

The circuit, shown in fig. 1, is fairly straightforward. An L-network is used for the input match and a pi-L network for the output. To operate in linear service the CM 10-12 must be forward biased. This is accomplished through the use of a CTC *byistor*. The byistor consists of a diode and a silicon resistor in one package which is coupled to the amplifier heatsink. The byistor thermally tracks the power amplifier and assures that problems with thermal runaway are minimized. A previous article describes this device in detail.[1]

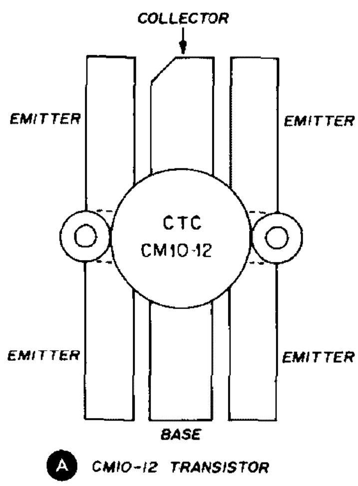


C1	2-20 pF mica trimmer (ARCO T51113-1)
C2,C3 C4,C5	4-40 pF mica trimmer (ARCO T21213-1)
C6	22 pF Underwood metal-clad mica
C7,C8 C9	33 pF Underwood metal-clad mica
C10,11	500 pF Underwood metal-clad mica
L1	0.175" (4.5mm) wide copper strip, 1.1" (28mm) long, including bend at input end; RFC1 attached at center
L2	0.175" (4.5mm) wide copper strip, 1.2" (30.5mm) long, including bend at output end; RFC2 and C3 attached at center
R1,R2	approximate total series resistance, 37 ohms (10 watts); adjust for correct byistor current (see text)
R3	adjust for correct idling current (6.8 ohms in parallel with 47 ohms, ½ watt, used in amplifier shown in photograph)
RFC1	no. 20 (0.8mm) wire, 1.5" (38mm) long
RFC2	2 turns no. 20 (0.8mm), airwound, 0.3" (7.5mm) diameter, 0.3" (7.5mm) long
RFC3	5 turns no. 20 (0.8mm) wound around 0.5" (12.5mm) Amidon T-50-6 toroid, in parallel with 15 ohm, ½ watt resistor

fig. 1. Solid-state 432-MHz linear amplifier. Do not substitute for the Underwood metal-clad mica capacitors; dipped micas or disc ceramics will not work.

construction

The amplifier is built on a 6x4.375 inch (15.2x11.1cm) finned aluminum heatsink (Thermalloy 6157). A smaller heatsink can be used, but I don't recommend it as parts placement becomes cramped and the CM10-12 power transistor will obviously run hotter. With the heatsink described a barely discernable heat rise is noticeable after several hours of operation.



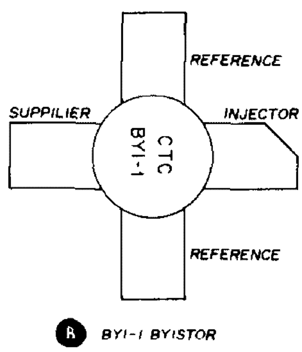


fig. 2. Lead layout for the CTC CM10-12 power transistor and BYI-1 byistor.

The end pieces are 1/8-inch (3mm) thick aluminum which are attached to the ends of the heatsink with 6-32 screws. The holes for the screws are drilled into the heatsink ends and then tapped.

The first step in building the amplifier is to prepare the printed-circuit board. The circuit board is standard double-clad 1/16 inch (1.5mm) thick fiberglass-epoxy board. The thickness is not really critical, but if the printed-circuit material is too thin the board may buckle when it is fastened to the heatsink. The CM10-12 power transistor is located at the exact center of the board. A 33/64 inch (0.516" or 13mm) hole is punched or drilled in the center of the board and the transistor is temporarily placed in the hole and used as a template to mark the locations for the two CM10-12 mounting screws. These holes are drilled (or punched) with a 17/64 inch (0.266" or 6.7mm) drill. The extra material is then filed away until the transistor mounting flange fits through the hole with about 0.015 inch (0.4mm) clearance all around (see **fig. 3**). Don't file away too much of the circuit board or it will be impossible to obtain short emitter leads — a must at uhf.

The 5/16 inch (8mm) hole for the byistor is next drilled 1.25 inch (32mm) from the center of the CM10-12. For best thermal tracking the byistor should be as close to the power transistor as is practical.

A good ground is a must in uhf power amplifiers and most problems with unstable amplifiers can be traced back to poor grounding. In this amplifier copper foil is soldered from the top to bottom ground planes around the entire outside edge of the board as shown in fig. 4. Similarly, short pieces of foil are soldered at the four edges of the CM10-12 mounting hole. Make sure

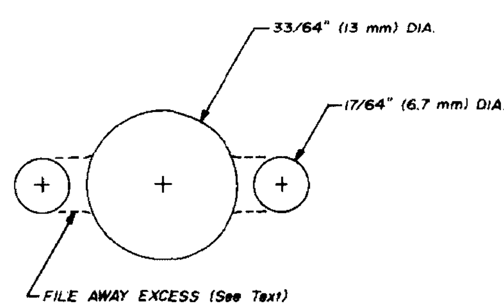


fig. 3. The CM10-12 transistor mounting hole is placed in the center of the circuit board (fig. 4).

these fit snugly or the transistor may not fit. These copper strips are about 0.125 inch (32mm) wide.

The board is then placed on the heatsink and secured with four 4-40 machine screws which are threaded into corresponding holes drilled and tapped into capacitor leads overlap one another (see **fig. 5**). This will provide the points for the base and collector connections.

The leads of the transistor are then trimmed as shown in the photograph. Put Silicone heatsink compound on the CM10-12 mounting flange and fasten

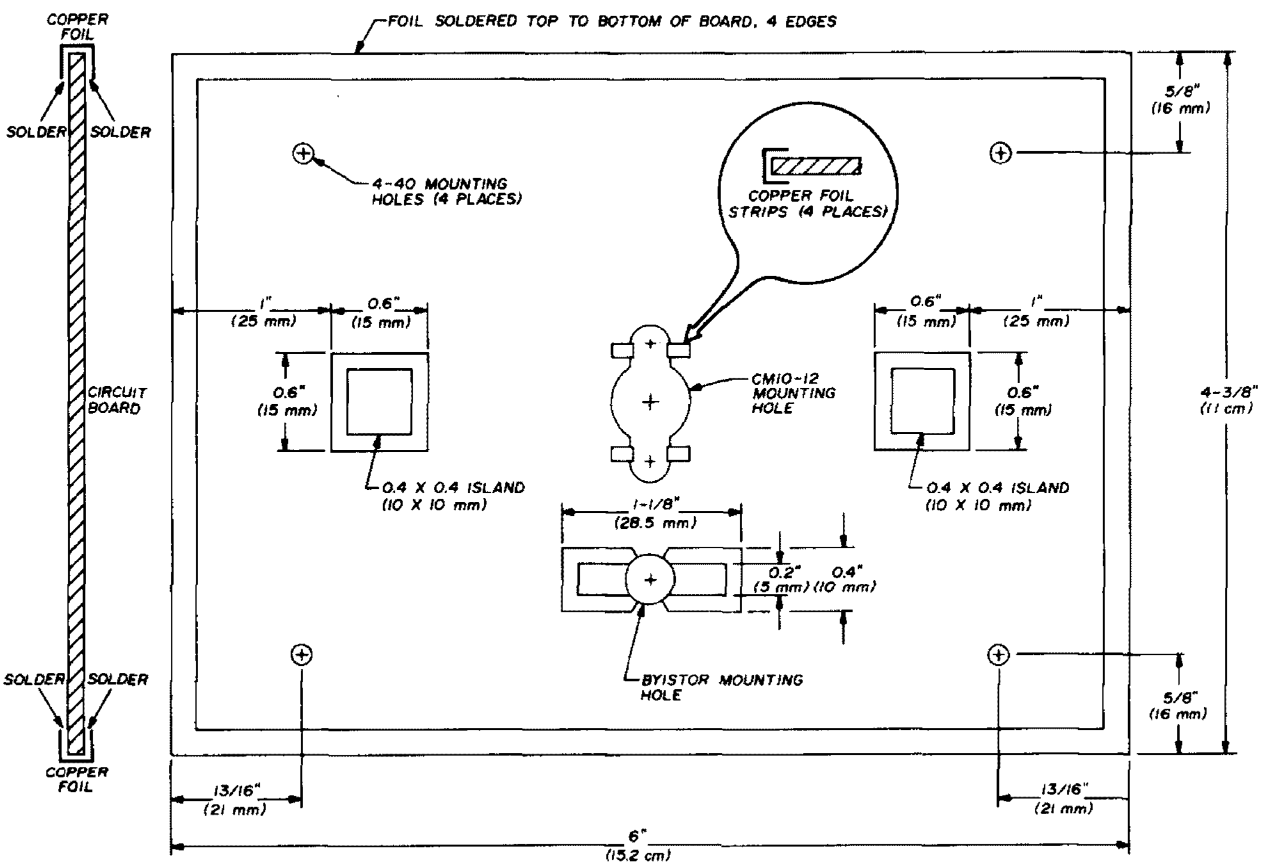


fig. 4. Layout of the circuit board used for the 432-MHz linear amplifier. Good grounding is provided by copper foil around outside edge of the board. Input and output islands and byistor circuit pads are cut out with an Xacto knife.

the heatsink. Temporarily remove the board and, with a milling bit or large diameter drill, prepare a space for the byistor mounting stud on the finned side of the heatsink. Be careful not to drill all the way through the heatsink.

The circuit board may now be secured to the heatsink with the 4-40 screws. Install the four Underwood capacitors* on the board as close as possible to the CM10-12 mounting holes. Solder the metal cases of the capacitors to the circuit board and let the the transistor to the heatsink. (*Do not* solder the transistor into circuit before securely fastening it to the heatsink with the two mounting screws – doing so may fracture the ceramic transistor case.) When the transistor leads have been soldered in place, small islands are cut out with an Xacto knife for the byistor supplier and injector and the transistor base and collector lines, L1

*Do not substitute for the Underwood metal-clad mica capacitors; dipped mica or disc ceramic capacitors will not work.

and L2. The remaining components are then installed as shown in the photographs.

Small turret terminals are used for the byistor power resistor connections and the 12-volt supply voltage, V_{CC}. It is suggested that the layout shown be followed if you have not had experience with uhf solid-state amplifiers.

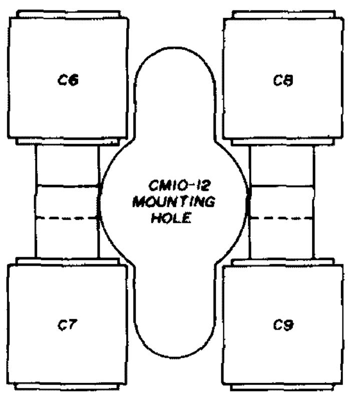


fig. 5. Installation of the Underwood mica capacitors, C6, C7, C8 and C9. Transistor base and collector connections are made at the overlap points.

tuneup

Do not operate the amplifier without resistor R3 connected; that will damage the power transistor.

Temporarily disconnect the collector dc feed choke, RFC2, gradually apply +12 volts to the byistor and make sure injector current is 300 and 350 mA. If this current level is not achieved, the values of resistors R1 and R2 must be adjusted until it is. Remove V_{CC} and reconnect the collector choke, RFC2.

Now install about 1 ohm at resistor R3, apply V_{CC} and check the total current drain. A milliammeter in series with the 12-volt supply line will read the byistor current plus the quiescent current of the power amplifier:

$$\text{Indicated current} = \begin{matrix}\text{byistor}\\ \text{current}\end{matrix} + \begin{matrix}\text{CM10-12}\\ \text{idling current}\end{matrix}$$

If the byistor current is 300 mA, for example, and the total indicated current is 350 mA, the idling current of the CM10-12 is 50 mA. Increase resistor R3 in one-half ohm steps until an idling current of 50 to 60 mA is reached. It may be necessary to *parallel two resistors* to obtain the correct idling current.

Basic tuneup of the amplifier can be done with a uhf vswr bridge, power meter and dummy load as shown in fig. 6. However, an accurate power meter and non-reactive 50-ohm load are required for proper circuit adjustment.

Apply V_{CC} and check to see that the power meter reads zero with no drive applied. This will indicate if the amplifier is oscillating. (No oscillation was detected in any of the four units I built.) Apply about 200 mW of drive and tune capacitors C3, C4 and C5 for maximum output. Tune C1 and C2 for minimum input vswr. Now increase drive to 600 mW and repeat the above process. The power meter should read approximately 10 watts.

If two-tone measurements are to be made and you're using an average-reading meter such as a Bird model 43, remember that the power meter indication will be one-half the actual rf output. For example, if the Bird wattmeter reads 5 watts, actual PEP output is 10 watts.

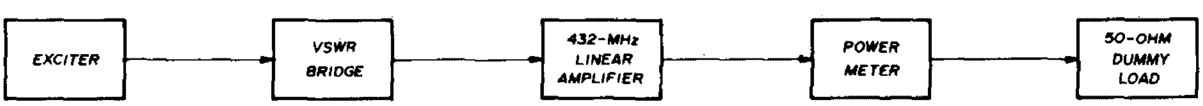
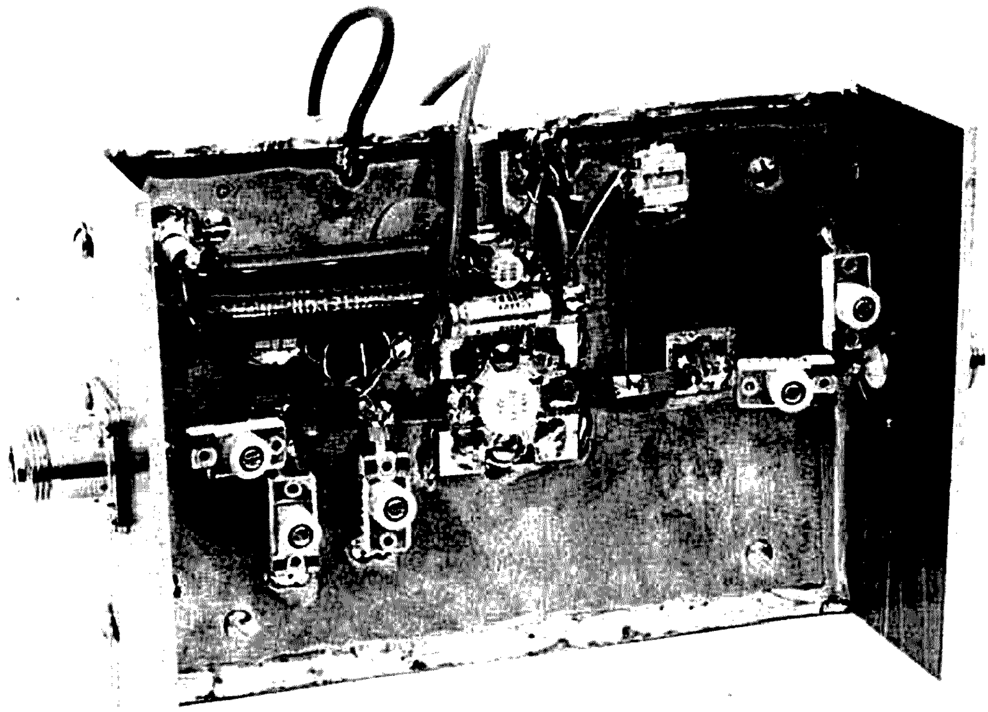


fig. 6. Equipment setup required for tuning up the 432-MHz amplifier. For correct adjustment the power meter must be accurate and the 50-ohm must be non-reactive.

operation

A well-regulated 12-volt supply is necessary for linear operation. Since current drain is less than 2.0 amps, a simple series regulator will work fine. Most current handbooks describe such circuits.

Since voice characteristics vary from one operator to another, if possible an oscilloscope should be used to check for flattopping. If a scope isn't available, talking collector current up to 500 to 600 mA will put you in the safe area (remember that the meter will read 300 mA high because of the byistor current). Make sure the amplifier works into a good load. Although the CM10-12 device will survive an infinite vswr, sustained operation into such loads will damage it.



Solid-state 432-MHz power amplifier built by WB6QXF. In this photograph output is at left, input on right. Large wirewound resistors are in the byistor circuit.

table 1. Performance of the 10-watt, 432-MHz solid-state linear power amplifier. Second harmonic is more than 35 dB down.

single tone		two tone	
V_{cc}	= +12.5 Vdc	V_{cc}	= +12.5 Vdc
P_{in}	= 15 Wdc	P_{in}	= 600 mW PEP
P_o	= 10 W	P_{out}	= 10 W PEP
I_c	= 1.2 A	I_c	= 750 mA
		IMD	= −29 dB at 10 W PEP

Every effort was made to make the construction of this amplifier as easy as possible. Inexpensive components were used throughout and the results have been very gratifying. The total cost of the amplifier is less than that of a comparable vacuum-tube unit and it is a lot less trouble.

reference

1. Robert Stein, W6NBI, "Solid-State Transmitting Converter for 144-MHz SSB," *ham radio*, February, 1974, page 6.

ham radio

adjustable voltage-regulator ICs

Circuit applications for the new Fairchild 78MG and 79MG adjustable positive and negative voltage-regulator ICs

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The 7800 series of fixed, three-terminal IC voltage regulators has greatly simplified the design of well regulated power supplies.[1] The only drawback of this series of ICs is that variable output voltages cannot be obtained without sacrificing performance or circuit simplicity. Fairchild Semiconductor has developed two new devices which fill the need for a low cost, adjustable voltage regulator: The 78MG positive regulator and the 79MG negative regulator.

Both of these IC regulators are rated for 500 mA output current, are protected against short-circuits and thermal overloads, and feature infinitely variable output voltages between V_{ref} and V_{in} (minus a few volts). Brief specifications are listed in table **1**.

positive voltage regulator

A basic, positive adjustable regulator using the 78MG IC is shown in **fig. 1**. The internal reference voltage for the 78MG is 5.0 volts, so this sets the lowest possible output voltage with this device. As is true with the 7800 series voltage-regulator ICs, a 0.33 μF capacitor is required from the input of the device to ground if the regulator is located more than a few inches from the power supply filter capacitor. A small capacitor from the output to ground will improve transient response.

The output voltage of this circuit is predicted by

$$V_{out} = V_{ref}\left(\frac{R1 + R2}{R2}\right)$$

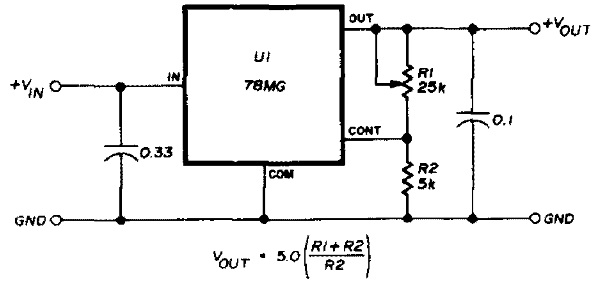

fig. 1. Basic, adjustable positive voltage regulator using the Fairchild 78MG. Unregulated input voltage must be 2 volts greater than desired maximum output.

where V_{ref} = 5.0 volts. If 1 mA is used as the control current, then R2 = 5k ohms and V_{out} = 0.001 (R1 + R2). With the values shown in **fig. 1** and V_{in} >32 volts, V_{out} is adjustable from +5 to +30 volts.

The 78MG may be used in higher current applications by using a series pass transistor, Q2, as shown in fig. 2.

table 1. Electrical parameters of the Fairchild 78MG and 79MG adjustable positive and negative voltage regulators.

78 MG positive regulator	
Input voltage, V_{in}	+40 volts max
Output voltage, V_{out}	+5 to +30 volts
Voltage reference, V_{ref}	+5.0 volts
Reference current, I_{ref}	1.0 μA
Dropout voltage	2 volts
Quiescent current	2.5 mA
Line regulation	1%
Load regulation	2%
Dissipation (internally limited)	≈ 6 watts
Thermal shutdown temperature	≈ 170°C
Thermal resistance, junction-case	8°C/watt
Thermal resistance, junction-ambient	70°C/watt
79MG negative regulator	
Input voltage, V_{in}	−40 volts max
Output voltage, V_{out}	−2 to −30 volts
Voltage reference, V_{ref}	−2.23 volts
Reference current, I_{ref}	0.3 μA
Dropout voltage	2 volts
Quiescent current	0.5 mA
Line regulation	1%
Load regulation	2%
Dissipation (internally limited)	≈ 6 watts
Thermal shutdown temperature	≈ 175°C
Thermal resistance, junction-case	8°C/watt
Thermal resistance, junction-ambient	70°C/watt

The short-circuit sensing resistor, R_{sc}, is equal to V_{be}/I_{sc}, or about 0.6 ohm for 1.5 amp output current. Transistor Q1 is another 2N6124 or equivalent. The sensing resistor, R_{sc}, turns Q1 on at high output current conditions and since $V_{ce(sat)}$ of Q1 is less than the voltage drop across R_{sc} plus the base-emitter voltage of Q2, the current through Q2 will decrease, protecting the transistor.

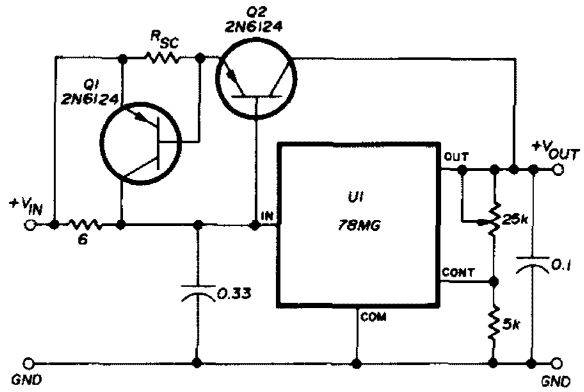

fig. 2. High-current, short-circuit protected adjustable positive voltage-regulator circuit using the Fairchild 78MG.

negative voltage regulator

A basic negative adjustable regulator is shown in **fig. 3**. The reference voltage for the 79MG is 2.23 volts and, as in the 78MG, this reference determines the minimum possible output voltage available from the device. Again, 1 mA should be selected for the control current; therefore, R2 is 2.2k ohms and V_{out} = 0.001 (R1 + R2). R1 may be a 25k ohm pot as was used in the circuit of **fig. 1**.

A dual tracking regulator with 500 mA output capability is shown in **fig. 4**. This circuit features dual 10 volt outputs with the use of only two ICs and seven external components.

packaging

Obviously, a four-terminal package is now required due to the addition of the control pin. **Fig. 5** shows the basic

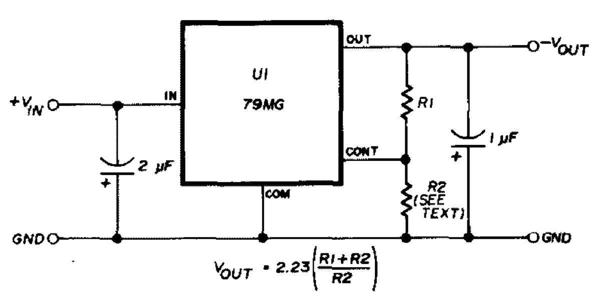

fig. 3. Basic, adjustable negative voltage regulator using the Fairchild 79MG. Unregulated input voltage must be 2 volts greater than desired maximum output.

configuration and connection diagram of both regulators, while **fig. 6** shows mounting techniques for two of the package options which are available. Package size and pin spacing are similar to the familiar mini-DIP package. For low power dissipation applications, the cooling wings may be bent upwards for natural convection cooling, or a heatsink may be added for higher power applications. Refer to **table 1** for thermal resistance and maximum junction temperature values for the package.

conclusion

These new adjustable voltage regulators fill the need for easy to use, high

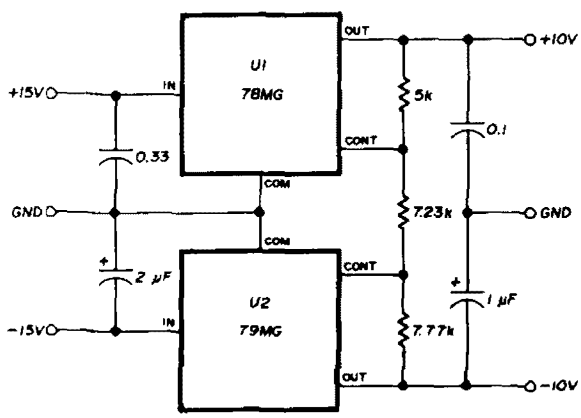

fig. 4. Dual tracking regulator for up to 500 mA output at ±10 volts.

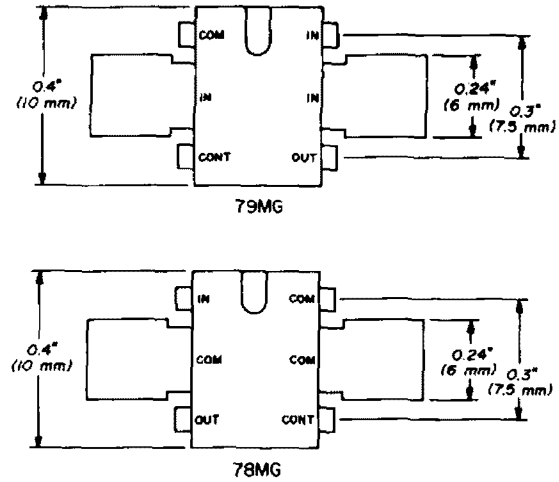

fig. 5. Connection diagrams for the 78MG and 79MG adjustable voltage-regulator ICs.

performance, minimum component count, variable-voltage regulators. Both electrically and thermally, the 78MG and the 79MG perform well in all medium-current applications and as drivers for higher current regulators. The 78MG and the 79MG and their related data sheets may be obtained from franchised Fairchild distributors.

reference

1. James Trulove, WB5EMI, "Three-Terminal Voltage-Regulator ICs," *ham radio*, December, 1973, page 26.

ham radio

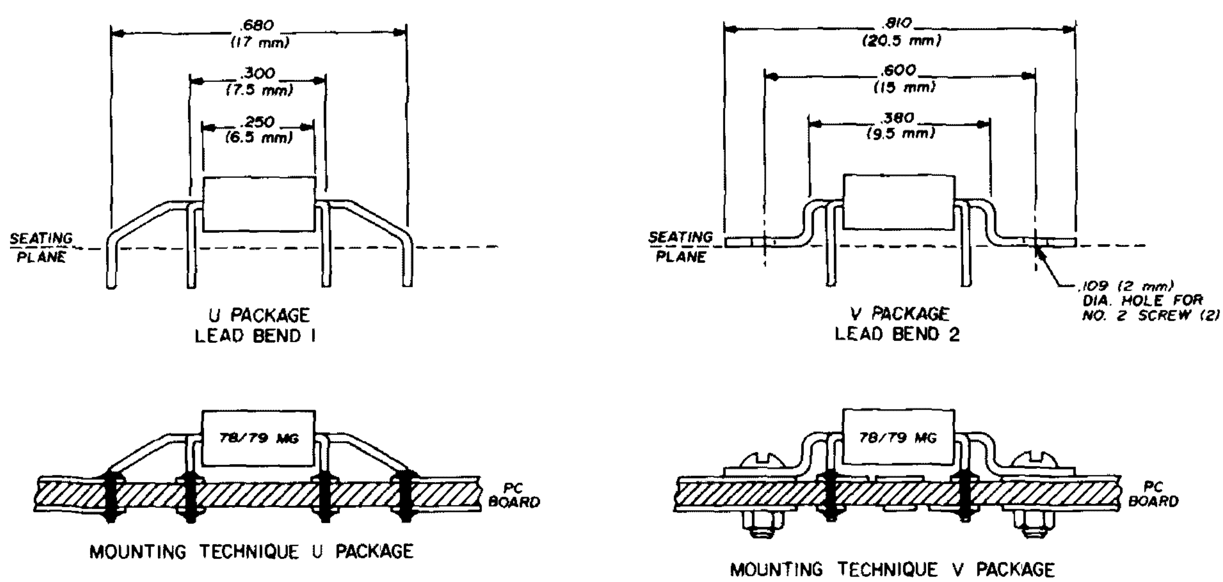

fig. 6. Dimensions and mounting techniques for the two package options which are available for the 78MG and 79MG adjustable voltage regulators.

calibrated electronic keyer time base

Description of a novel electronic time base that provides direct readout of keyer speed

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It is desirable to be able to set the speed of an electronic keyer before you start sending. This is difficult if the keyer has a single uncalibrated knob and the variable resistor it controls is not linear and gives a large change in speed for a small rotation. Three solutions to this minor problem will be discussed in this article: speed selection with toggle switches or a thumbwheel switch, a digital counter, and a simple analog frequency meter which will display keyer speed on a meter. The last two systems require the time-base oscillator to run continuously. However, by running the oscillator at a high frequency and dividing down, the normal disadvantages of a continuously *running* oscillator can be avoided.

switch selection

It is well known that the time delay of an RC circuit is proportional to the RC product. The frequency of an RC oscillator, therefore, is proportional to 1/RC, or G/C where G is the conductance (1/R) of the resistor. When resistors are connected in parallel the conductances add. Resistors can easily be switched in parallel with toggle switches or thumbwheel switches, and when this is done the frequencies associated with the individual resistors add directly.

In the keyer time base shown in fig. 1, a 560k resistor (R5) is permanently wired in to give a speed of 5 wpm. Another 560k, 270k or 150k resistor can be switched in to add 5, 10 or 20 wpm to the speed. With three switches any speed from 5 to 40 wpm can be selected in steps of 5 wpm. The 1000-ohm vernier adjustment pot, R1, can be mounted on the front panel if continuous speed adjustment is desired. This pot can also be screwdriver adjusted (or even omitted if the timing capacitor is carefully selected).

Fig. 2 shows how a single thumbwheel switch can be used to provide keyer speeds of 5 through 45 wpm. **Fig. 3** shows how a two-section thumbwheel

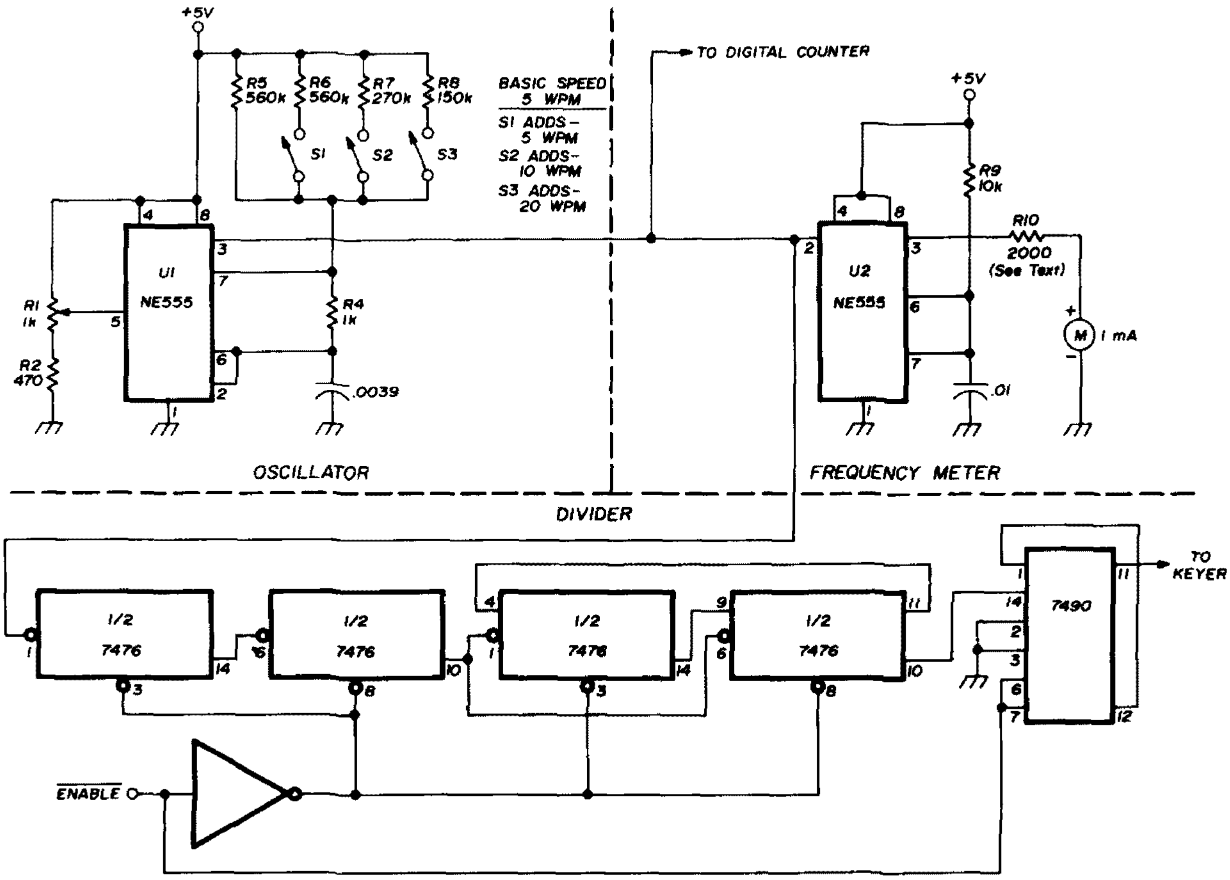


fig. 1. Electronic keyer time base with switch selection of speed. Analog frequency meter indicates keyer speed directly.

switch can be wired for speeds of 1 through 99 wpm in 1 wpm steps. The principle of paralleling resistors is very simple and can be applied to any keyer whose speed is determined by a single resistor and capacitor.

digital speed readout

A formula in the ARRL *Radio Amateur's Handbook* states that the speed of an electronic keyer is 1.2 times the clock frequency in Hz. In the time base shown in **fig.** 1 the frequency of the basic NE555 oscillator is 100 times the keyer speed. The keyer clock is obtained by dividing the oscillator frequency by 120. For a 24 wpm keyer speed, for example, the oscillator runs at 2400 Hz. This can easily be read on a digital counter. If the counter is set to its 100 Hz range (dropping the tens and units digits), the speed can be read directly. The time-base divider would supply a clock frequency of 20 Hz, the correct clock frequency for 24 wpm keying.

Normally an electronic key with a continuously running oscillator does not have a good "feel" because you have to wait up to one-half the length of a dot for the keyer to start after pressing the paddle. With the time base shown in **fig. 1**, however, you only have to wait 1/240 the length of a dot, which is hardly noticeable (only 2 milliseconds

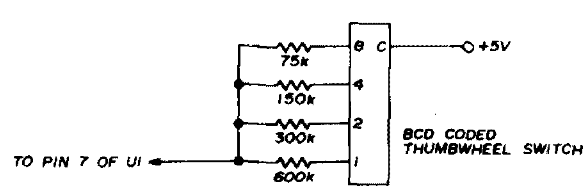
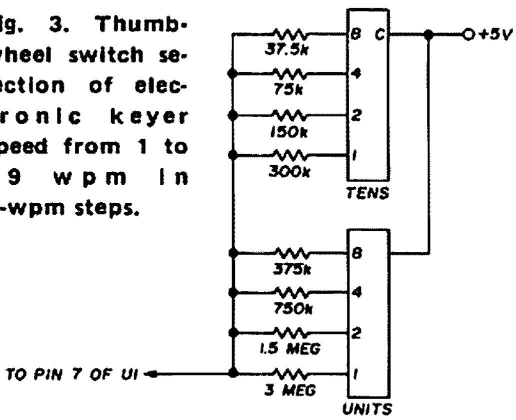


fig. 2. Using a thumbwheel switch to set electronic keyer speed. This arrangement provides speeds of 5 to 45 wpm in 5-wpm steps.

at 5 wpm). This is because when the keyer is not enabled the dividers are poised with all outputs high, ready to give a negative transition as soon as the next oscillator pulse comes along after the keyer is enabled. This system has all the advantages of a continuously running oscillator with none of the disadvantages.

fig. 3. Thumbwheel switch selection of electronic keyer speed from 1 to 99 wpm in 1-wpm steps.



The signal marked "to keyer" should go to a JK flip-flop such as the 7473 or 7476 which responds to negative transitions. The input to the inverter should be high when not sending and low when sending or self-completing a character.

keyer speed meter

It's doubtful that anyone would build a digital counter especially for his electronic keyer, and switching another counter to the keyer is not as convenient as simply looking at a panel meter which indicates keyer speed. The simple analog frequency meter shown in **fig. 1** uses few components and has all the accuracy that is needed. The meter is a 0-1 milliammeter with a 0-5 scale such as the Lafayette 99F26270. Resistor R10 should be trimmed to give a full-scale reading at 50 wpm using a digital counter to set the speed to 50 wpm (oscillator at 5000 Hz). The meter in my keyer, calibrated at 50 wpm, agrees with my digital counter over the entire scale.

ham radio

dc latch circuit

Modern dc latch for transmitter PTT control uses cmos circuits for low current drain

Bill Lambing, WØLPQ, 1480 25th Street, Marion, Iowa 52302

A dc latch circuit or flip-flop is a handy circuit to have around if you operate PTT but don't want to hold down the microphone switch. One circuit described by W2EEY[1] used a form of latch for PTT-PTL (push-to-talk, push-to-listen) control of a radio system.

Fig. 1 shows a fairly typical latch circuit of the 1960s which used four bipolar transistors and one unijunction *to drive a relay circuit.* Applying a +12 volt pulse to diode CR1 caused the circuit to be *set*. Applying a +12 volt pulse to diode CR2 *resets* the circuit. However, during the set cycle the 180-ohm resistor in Q2's emitter circuit dissipates almost 800 mW of power. Not counting the relay, this circuit requires almost 100 mA (1.2 watt) to function – with modern cmos circuits this excessive power drain can be reduced by a factor of 200.

updated latch

The latch circuit shown in **fig.** 2 is made for *today. Disregarding* the transistors Q3, Q4 and Q5 and the two relays for a moment, when the push-button switch, S1, is depressed, the circuit draws slightly more than 500 microamps. When switch S1 is not activated current drain is in the region of 100 μA. This very low power requirement is one of the advantages of using modern cmos ICs. This is not to say that cmos doesn't have its problems. It does: handling, static discharge and drive ability, but for ultra low power consumption cmos is the answer.

In the circuit of **fig.** 2 integrated circuit U2A is a dual type-D cmos flip-flop. When the incoming clock signal makes a transition from zero to 1, the signal on the data input (D) is gated through to the Q output. When switch S1 is depressed, the input to the cmos inverter, U1A, goes low and its output goes high. The RC network consisting of the 100k resistor and 0.01 μF capacitor add roll-off to the leading edge of the zero to 1 transition, eliminating any contact bounce caused by S1. It has been found by experimentation that the RC time constant should approximate (or be slightly longer) than the contact bounce of your particular switch; more about this later.

During the zero to 1 transition, if the Q output of U2A is high, $\bar{Q}$ by definition is low. Since this low is also present on the data input as the clock makes the transition from zero to 1, Q goes to zero and $\bar{Q}$ goes to logic 1. At the next zero to 1 clock transition the output states will change.

When the Q output of U2A is low, it is inverted in U1B, turning on transistor Q2 and enabling relay K1. The $\bar{Q}$ output, which is high, is inverted in U1C, turning off Q1 and relay K2. The 4700-ohm resistors in series with the inverter outputs are determined by the

transistor collector loads. Do not try to use a higher value or the transistor will not turn on. A lower value will most likely burn out the cmos inverter. Note that the series resistor R2 is 10k because Q3's collector load is greater than the loads seen by Q1 or Q2.

The main function of transistors Q3, Q4 and Q5 is as a power switch. When transistor Q3 is turned by a logic 1 at the $\bar{Q}$ output of U2A, it allows +24 volts (or what is on the Q5 emitter) to be seen at Q5's collector. Several hundred milliamperes of current can be switched in this way using a 2N2222A. More current can be switched by using a Darlington connection at Q5.

When working with this circuit just remember that a logic zero at the base of Q3 turns on transistors Q4 and Q5. A logic 1 turns on Q3, pulling the base of Q4 to ground, which shuts off power switch Q5. If you don't mind relay noise, you can use a relay. However, if you like silent operation and no wasted relay power, use the solid-state arrangement.

If you consider transistor Q3 by itself (forgetting Q4, Q5 and the three 10k resistors for a moment), you can tie the collector of Q3 to your speaker output for muting. This arrangement can't be used, however, if your speaker is driven directly by a complementary bipolar or IC output device; it can only be used where an output transformer is used (no dc component). Otherwise connect the Q3's collector across the volume control for an audio signal without a dc component.

cmos levels

The graph of fig. 3 shows a large intermediate region in the operating characteristics of cmos logic that expands as the supply voltage, V_{DD}, is increased. This is where cmos really shines — noise immunity. The maximum acceptable input level for a cmos device in a low-level input state, V_{IL}, is 30 per cent of V_{DD}, or 4.5 volts when V_{DD} = 15 volts. This means that a logic zero can be from 0.01 to 4.5 volts.

At the other end of the scale, the minimum acceptable input level for a cmos device in a high-level input state, V_{IH}, is 70 per cent of V_{DD} or 10.5 volts when V_{DD} = 15 volts. This means that a logic 1 can be 10.5 to 14.99 volts. The intermediate region (or noise margin) is from 30 to 70 per cent of V_{DD}.

Referring back to fig. 2, to effectively combat contact bounce the RC time constant at one time constant (63.2%) should approximate or be slightly longer than the switch's contact

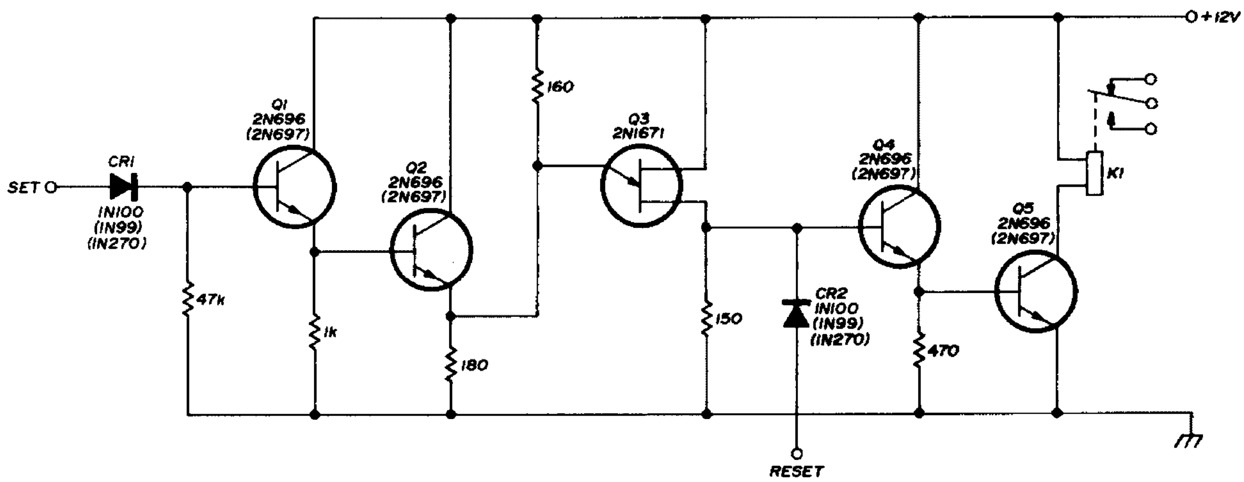


fig. 1. Dc latch circuit using bipolar transistors requires up to 1.2 watt of power. In this circuit a +12 volt pulse at set input latches relay K1; +12 volt pulse at reset input resets K1.

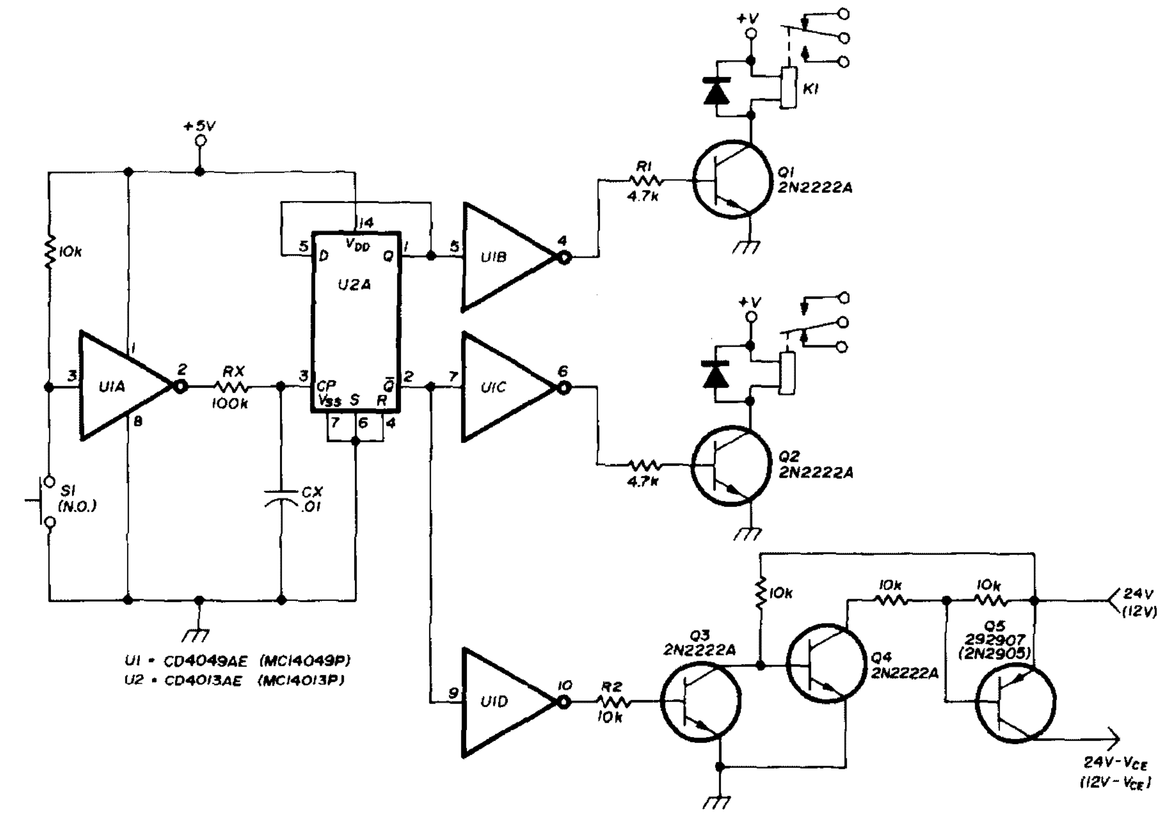


fig. 2. Modern dc latch uses cmos ICs for minimum current drain. Either relay switching (Q1, K1 or Q2, K2) or solid-state switching (Q3, Q4 and Q5) may be used.

bounce, and should approximate 70 per cent of the supply voltage, V_{DD}.

When using cmos devices be sure you look at the manufacturers' specifications.[2] If you are not using an input, tie it to ground or to V_{DD} whichever is appropriate. Do not use cmos NAND or NOR gates to drive transistors – inverting or non-inverting buffers are designed to do this job. Be careful when handling cmos devices as the static burnout problem is very severe. In addition, do not use plastic bags or polyethylene snow for packaging cmos devices – this stuff has zapped more than one device.

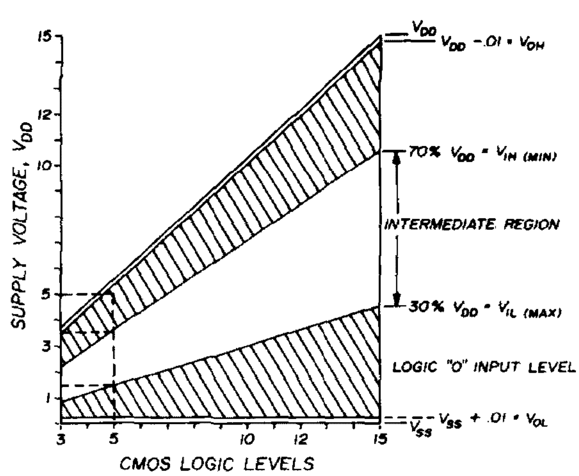


fig. 3. Graph of cmos logic levels shows how intermediate region with high noise immunity expands as the supply voltage is increased. Dashed lines show operation with +5 volt supply (see text).

Reference 3 has much more information on the use of cmos devices which should enable experimentally inclined amateurs to learn much more about these very useful devices. Just obey the rules and you can gain as much as a 1000:1 reduction in circuit current drain.

references

1. John Schultz, W2EEY, "Solid-State Transmitter Switching," *ham radio*, June, 1968, page 44.
2. *COS/MOS Digital Integrated Circuits*, RCA Solid-State Data Book SSD-203B, RCA Solid State, Somerville, New Jersey 08876.
3. M. Stiglianese, "Interface CMOS Logic with Switches," *Electronic Design*, August 16, 1974, page 80.

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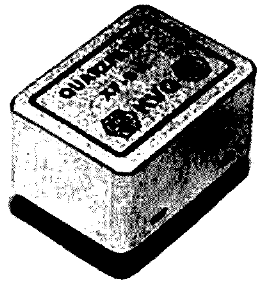

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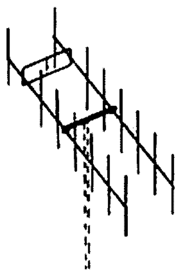

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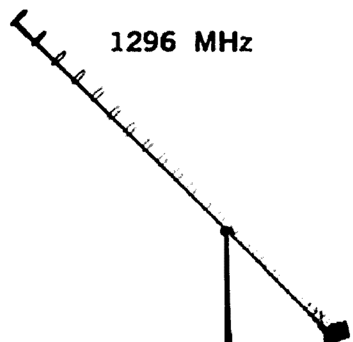

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More Details? CHECK—OFF Page 94

fet-controlled charger

for small nicad batteries

A novel use of the fet produces this simple constant-current nicad charger

G. Kent Shubert, WAØJYK, 1308 Leeview Drive, Olathe, Kansas 66061

Rechargeable nickel-cadmium batteries have become very popular over the last few years, not only for hand-held fm transceivers, but also for powering test equipment. Several ideas for inexpensive chargers have been tried, but maintaining a constant rate of charge has been a problem.[1]

The Regency HRT-2 uses the Regency MA-50, Eveready N64, Gould CD64, or the Alexander R64 rechargeable nicad battery, which is representative of the batteries used in most handheld units. Regency recommends a charge rate of 50 mA and a trickle or "float" rate of 15 mA. It is detrimental to the nicad if the voltage across the battery rises too high. This particular battery is fully charged at 14.4 volts, so a 15-volt zener diode was chosen to provide over-voltage protection.

constant current charger

Junction field-effect transistors can be used as constant-current sources simply by shorting the gate to the source. The current which results is the I_{dss} rating given in the data sheets. Type 2N3819 fets were used in the circuit in fig. 1 simply because there was an abundance of them on hand. Practically any n-channel junction fet will work, but only devices that have an I_{dss} of 8 to 15 mA should be used. There are special power fets available, but most of the inexpensive plastic devices will strain to dissipate a quarter of a watt at room temperature — so don't crowd them. The six fets used were actually graded by connecting them as shown in fig. 2 and grouped to supply the 15 mA or 50 mA as selected by the switch.

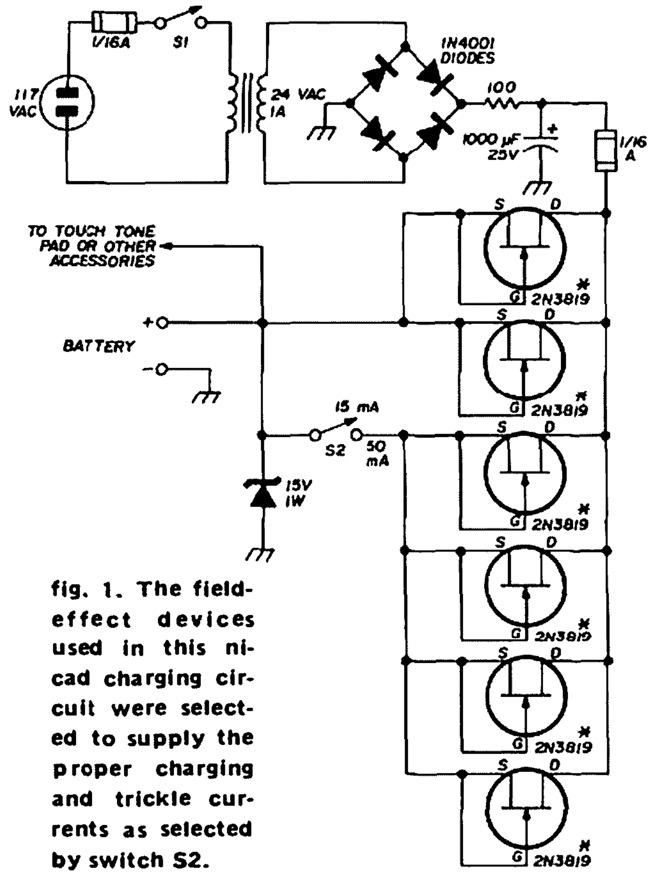


fig. 1. The field-effect devices used in this ni-cad charging circuit were selected to supply the proper charging and trickle currents as selected by switch S2.

The photograph shows a few other frills that were added to the basic charger to make life on two meters a little more enjoyable. The hand-held portable becomes a low-power base station when placed in the charger. The HRT-2 has both antenna and battery terminals in the base, so placing the unit on the charger connects the external antenna and places the battery on charge. The external microphone and PTT keying is connected to the top of the HRT-2 to enable the hand microphone and the Touch-Tone encoder to function. The ac power is switched by a microswitch mounted so the weight of the transceiver turns it on. The unit is held firmly to the charger with a black elastic garter, a difficult item to locate nowadays.

As an added bonus, the unit can be used far from ac power lines just for the Touch-Tone encoder function. It is necessary to use an external antenna, but at least the charger serves as an adapter from the mini-phone plug to the more popular UHF or BNC connectors.

The hookup of the Touch-Tone pad is standard with fm operators so no details are included here. The 1/16th amp fuse is not necessary but is added life insurance against a semiconductor failure.

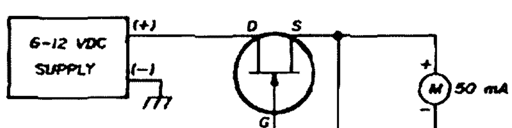


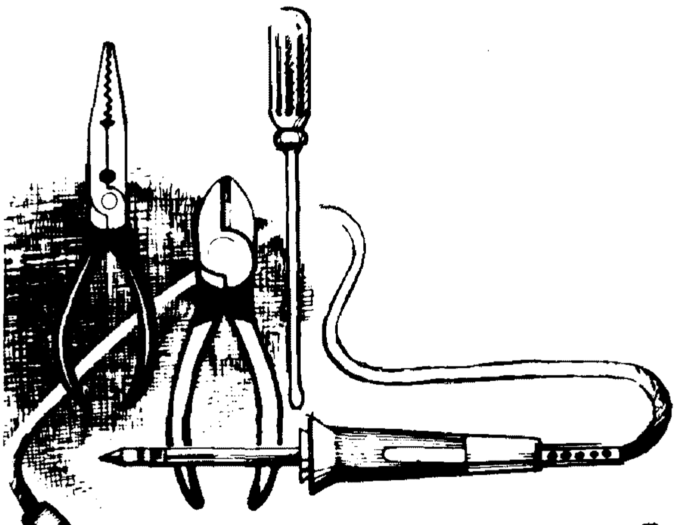
fig. 2. Setup for testing fets for the drain-to-source current, I_{dss}, measured with the gate shorted to the source.

reference

1. R. D. Shriner, WAØUZO, "Charging Nickel-Cadmium Walkie-Talkie Batteries," *QST*, August, 1973, page 44.

ham radio

The Regency HRT-2 sits atop the charger overlooking the ever popular Touch-Tone pad.



circuits and techniques

ed noll, W3FQJ

QRP fet transmitter

The **Siliconix 2N3970** switching fet performs well in low-powered transmitter circuits such as crystal oscillators, modulators, rf amplifiers and frequency multipliers. Circuit simplicity is an fet advantage and many circuits are identical to vacuum-tube arrangements except that no filament power is required. The two-stage crystal oscillator and amplifier, **fig. 1**, requires approximately 500 mW dc input to the final and can be operated from two 12-volt lantern batteries in series. At W3FQJ it operates from the solar power supply detailed in the November, 1974, issue of *ham radio*.[1] A 12-volt motorcycle battery has also been added to the installation, providing 24-volt capability.

The oscillator is a Pierce-type and requires no resonant output circuit. The signal is capacitively coupled to the gate of the amplifier with an rf choke serving as a means of applying the drain supply voltage. A 1-mA meter can be connected across the low-value gate-circuit resistor to provide an indication of gate current and, therefore, the strength of the oscillator signal arriving at the gate. The meter can be connected and disconnected without any influence on the operation of the transmitter because the value of R4 is very low in comparison to the value of the gate resistor, R3. The

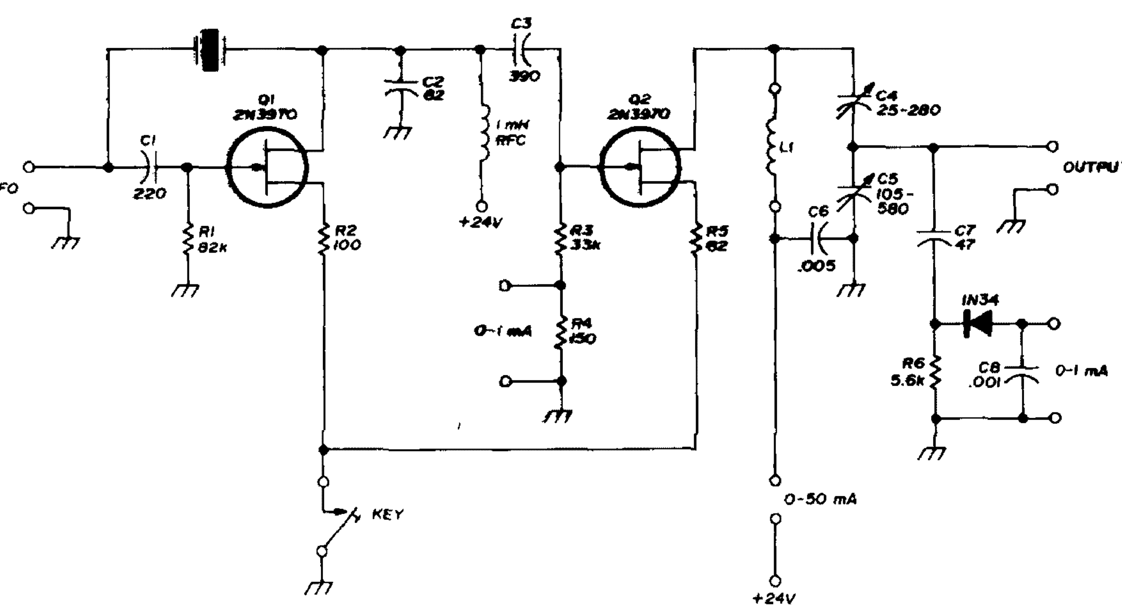


fig. 1. Two-stage fet QRP transmitter uses crystal oscillator stage, Q1, and power amplifier, Q2. Dc power input is about 500 mW. A vfo may be used if desired. For 80 meters L1 is 50 close-spaced turns no. 24 on 13/16" (20mm) toroid core.

low-value capacitor C2 aids crystal starting when operating at low frequencies. Parts values and information are given for operation on the 80-meter band.

The resonant circuit of the rf amplifier consists of a toroid coil and two series-connected trimmer capacitors. These capacitors are used both for tuning and for obtaining an impedance match to a low-impedance load. Capacitor C4 influences primarily the resonant tuning; capacitor C5 is for the impedance match. The amplifier drain current can be measured with a 50-mA meter or an appropriate vom current scale. The product of drain current and supply voltage represents the dc input power to the amplifier. As in vacuum-tube practice, when the resonant circuit is tuned through the resonant point, there is a dip in drain current.

A 1N34 diode and resistor-capacitor filter are used as an rf output indicator. The dc current indicated on the 1-mA meter is a relative measure of the level of the rf output voltage. This meter can also be connected and disconnected without affecting the output power level.

The simple two-stage QRP transmitter is mounted on a Vector board, fig. 2. This is micro-vector board type 84P44-062; the 0.042-inch (1-mm) hole size and 0.1 inch (2.5mm) spacing are ideal for mounting transistor and IC sockets. The Vector T42-1 micro-clips can be inserted conveniently into the holes to provide terminals. This method of construction will be used throughout the *Expro* projects. Binding posts are convenient for making tests and interconnections. Stick-on protector pads (available at hardware stores) provide support and permit the bulk of the wiring to be done underneath the Vector board.

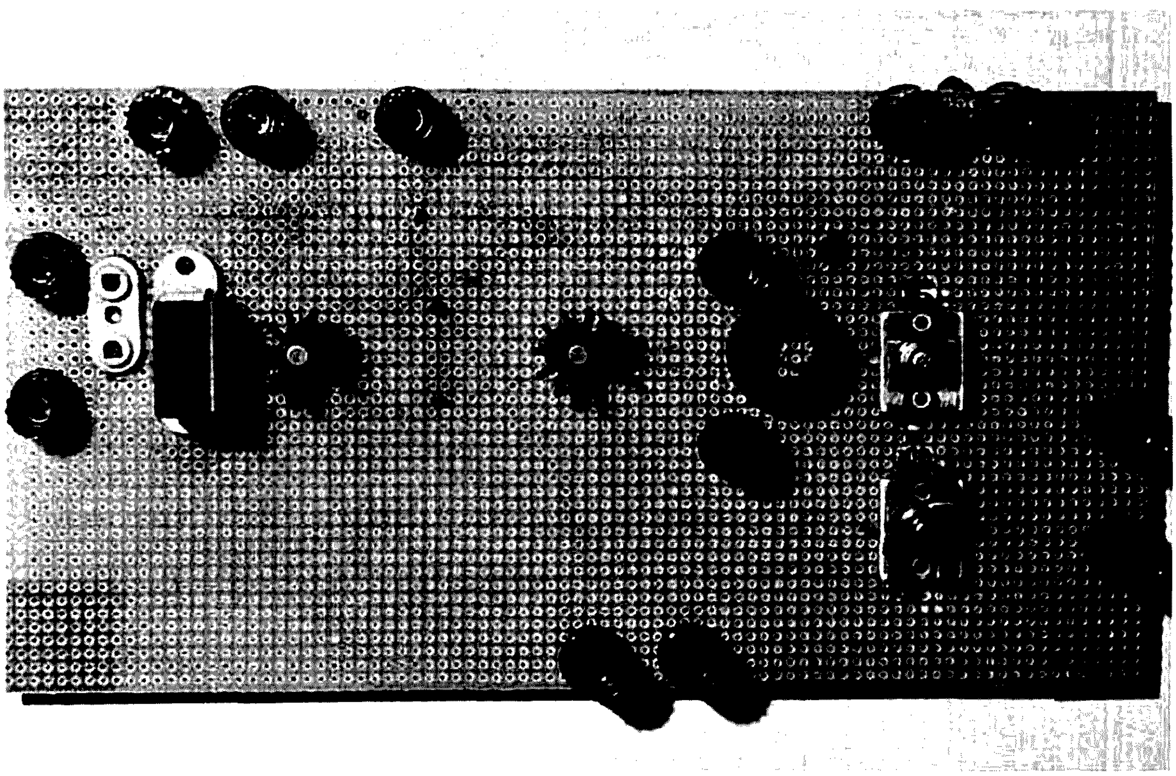


fig. 2. Two stage fet QRP transmitter is built on piece of Vector perfboard as shown here.

The toroid core is the 13/16-inch (20-mm) type. The winding consists of

itors are mounted near the coil so they can be easily adjusted while observing the readings on the output indicator.

You may wish to drive the two-stage transmitter with a variable-frequency oscillator connected to the vfo input binding posts, fig. 1. For this mode of operation you need only remove the crystal from its socket.

fig. 3. For portable use you may want to use a rechargeable Gel/Cel battery made by Globe Battery, Milwaukee, Wisconsin.

50 close-spaced turns of number-24 enameled copper wire. There are two binding posts positioned on each side of the toroid. Later you may wish to operate on other bands and they provide an easy means of changing coils for multiband operation. The two trimmer capac-

tuneup

To check out the transmitter, first insert the crystal and the two fets. Arrange the supply voltage so as to apply power, initially, to the oscillator only. Connect the 1-mA meter across the gate resistor R4.

Turn on the oscillator stage. Note that there is an indication on the meter. This indicates that the oscillator is operating and there is adequate drive to the amplifier, enough to draw gate current. Meter reading is low and approximately 0.1 mA. Note that if the crystal is removed from its socket, the current reading falls to zero. Also, if the amplifier fet is removed from its socket the meter reading falls to zero.

Insert the 50 mA meter in the supply line to the drain of the amplifier. Adjust capacitor C5 for near maximum setting. Apply power to both stages. Now tune capacitor C4 through its range. Note the

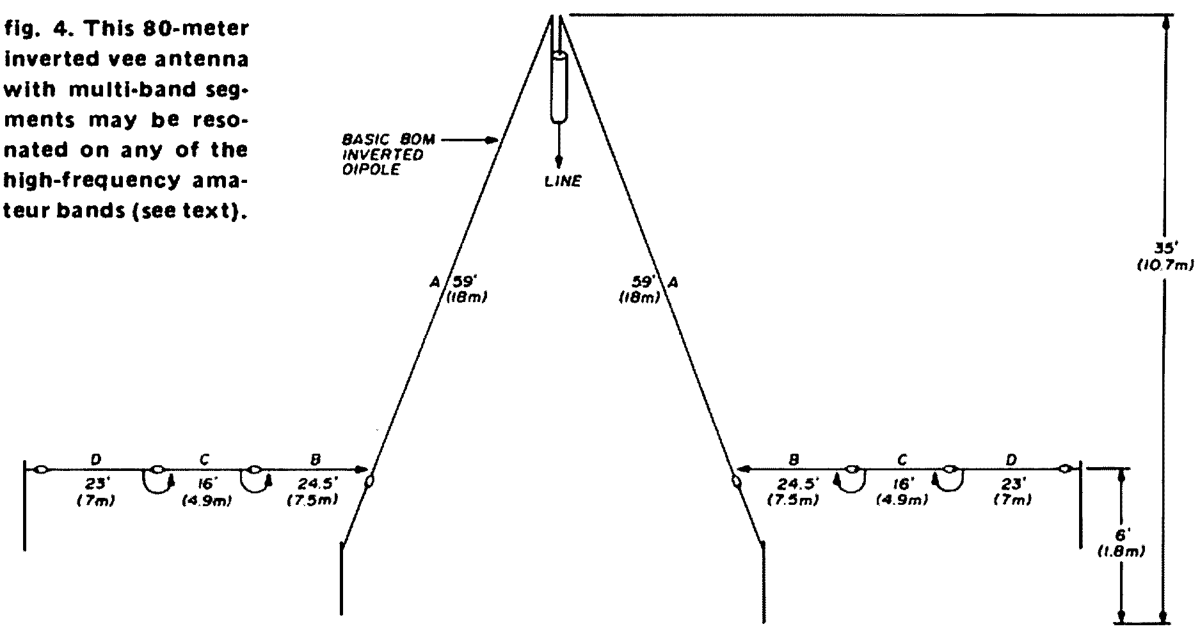


fig. 4. This 80-meter inverted vee antenna with multi-band segments may be resonated on any of the high-frequency amateur bands (see text).

dip in the drain current as the output circuit is tuned through resonance.

Transfer the 1-mA meter to the indicator circuit (across capacitor C8), connect a 68-ohm resistor across the amplifier output, and turn on the transmitter. Adjust capacitor C4 for maximum meter reading. Now adjust capacitor C5 for best output. Jockey back and forth between C4 and C5 until maximum output is obtained.

Jot down the output meter readings and the drain current reading, and calculate the dc power input to the amplifier

$$P_{IN} = V_{DD} \ I_D$$

Typical dc input power is 480 milliwatts (24 V x 20 mA).

Now connect an oscilloscope across the output. Note the good quality of the generated 80-meter sinewave. Key the transmitter, noting the influence on the oscilloscope pattern and the drain and output current meter readings. Tune the transmitter in on your receiver. Check out the keying quality of the CW signal.

Remove the 68-ohm resistor which is across the output, disconnect the oscilloscope, and connect your 80-meter dipole antenna across the output. Retune

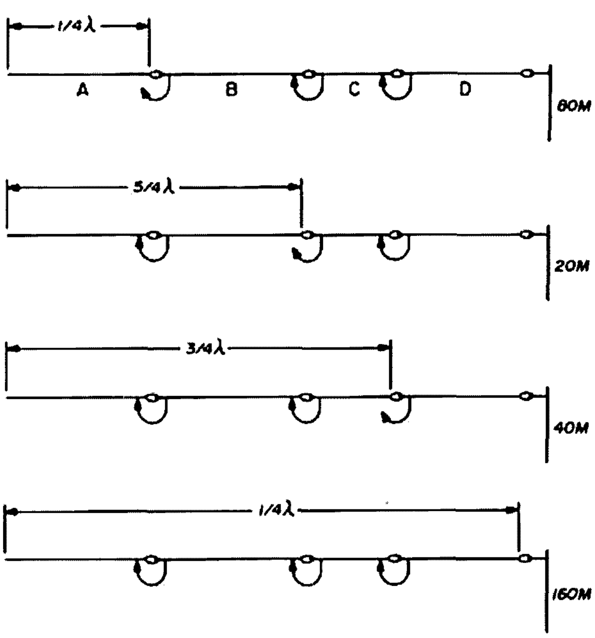


fig. 5. How to use the basic antenna system of fig. 4 on four bands by using jumpers.

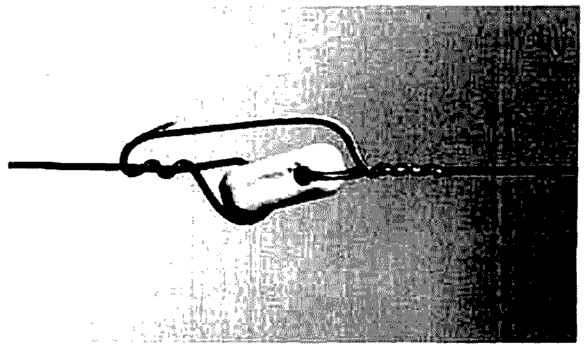


fig. 6. Installing a simple jumper wire across the antenna insulator to adjust resonance.

the transmitter. If your antenna system matches properly there should be very little change in drain current or the output meter reading. If your antenna is not resonant to exactly the crystal frequency the meter readings might not be the same. You are now ready to operate your fet QRP transmitter. Initial W3FQJ contacts with this little rig were with W2UEZ and W2UUV/1.

The rig can also be operated portable using the Globe Gel/Cel 4.5AH 12-volt battery. This convenient battery can be supplied with its own charger or it can be charged from a small solar energy converter, fig. 3.

single antenna — four bands

In my solid-state and QRP experiments I needed a single good performing, one-transmission-line test antenna for the 20-, 40-, 80- and 160-meter bands. In addition, it should be possible to resonate the antenna to any frequency in any band. For good performance there should be a current loop at the top of the antenna for each band. For convenience it is helpful to make all resonant frequency and band changes from the ground level and without letting a mast down or putting it back up. A bit of walking to make changes was welcome rather than frowned upon.

The Inverted-Vee antenna was selected because current maxima could be positioned at the apex for each band by proper selection of leg length. At the

same time all changes in resonant leg length could be made from ground level. This is accomplished by making each leg length some odd multiple of an electrical quarter wavelength, reflecting a low impedance to the feedpoint at the apex. The final antenna operated as an inverted dipole on 80 and 160; a 3/2λ Inverted-Vee on 40; and 5/2λ on 20.

A general plan of the antenna is shown in **fig. 4**. The 80-meter segment is a conventional inverted dipole with its apex about 35 feet (11m) up at W3FQJ, with wire ends reaching down to 4 to 5 feet (1 to 1.5m) above ground level. From these accessible ends the legs of the antenna span out horizontally at the same level.

Segment B, approximately 25-feet (7.6m) long, when added to segment A with a jumper in each leg sets up the 5/2λ antenna on 20 meters. Additional 15-foot (4.5m) segments jumpered onto leg ends establishes a 3/2λ on 40. Finally, about 25 additional feet (7.6m) provide a half-wavelength antenna on 160 meters. The legs do not necessarily have to run straight away. When necessary they can be tilted away from the plane of the Inverted-Vee by as much as 40 to 60° to permit accommodation to the mounting site.

The arrangement of each leg of this antenna is shown in **fig. 5**, showing the jumper open or closed plan for four-band operation. All of this can be done conveniently from ground level. The photograph of **fig. 6** shows how a jumper is closed across a standard ceramic insulator.

table 1. Free-space dimensions (f in MHz).

1/2 wavelength	492/f (feet)	150/f (meters)
3/2 wavelength	1496/f (feet)	450/f (meters)
5/2 wavelength	2460/f (feet)	750/f (meters)
7/2 wavelength	3444/f (feet)	1050/f (meters)

Free-space dimensions for a sequence of odd quarter-wavelength segments is shown in **table 1**. These free-space lengths must be shortened to obtain an electrical resonance with a wire antenna. The following formulas are normally used to find the length of a quarter-wavelength dipole leg:

$$\text{leg length (feet)} = \frac{234}{f_{MHz}}$$

$$\text{leg length (meters)} = \frac{71.3}{f_{MHz}}$$

This works for most amateur bands. However, when building a 160-meter antenna recently I found the leg length was more closely given by $228/f_{MHz}$ (feet) or $69.5/f_{MHz}$ (meters), possibly showing the close-to-ground influence.

table 2. Design equations and resonant points for inverted-vee antenna shown in fig. 4.

band	antenna	aquations		measured resonance
160 M	1/4 wavelength	228/MHz (feet)	69.5/MHz (meters)	1850 kHz
80 M	1/4 wavelength	234/MHz (feet)	71.3/MHz (meters)	3930 kHz
40 M	3/4 wavelength	725/MHz (feet)	221/MHz (meters)	7290 kHz
20 M	5/4 wavelength	1210/MHz (feet)	369/MHz (meters)	14340 kHz

By experiment I have found that the leg-length equations for 3/4 wavelength are $752/f_{MHz}$ (feet) and $221/f_{MHz}$ (meters); for 5/4 wavelength the equations are $1210/f_{MHz}$ (feet) and $369/f_{MHz}$ (meters). These equations should get you into each of the desired bands. However, some length ad-

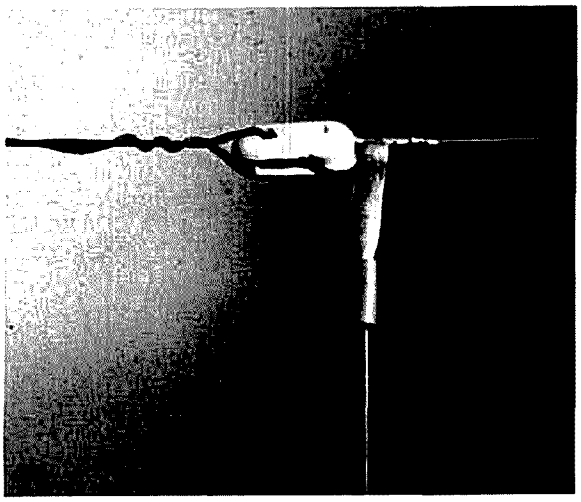


fig. 7. Resonating to a spot frequency in any band may be accomplished with short clip-on sections.

justments will undoubtedly be necessary for resonance at your preferred frequency as variables inevitably creep into any antenna installation.

Resonating to a spot frequency within any one band can be handled with clip-on sections, fig. 7. In using the idea of the clip-on section, dimension the antenna segments to the high end of each band. For example, the 80-meter inverted dipole is resonated near 3.95 MHz. Two clip-on sections of proper length can then be used to resonate the dipole to any lower frequency in the same band. For the case in point clip-on lengths of 4 feet (1.2 meter) tune the antenna to resonance at 3.6 MHz.

The dimensions given in fig. 4 are the final practical values. Data for them are given in table 2. Of course, you may wish to cut the antenna segments for resonance at the center of the phone segment of each of the bands. If you decide to do this, a single pair of clip-ons for each band can then be used to lower antenna resonance into the CW band.

reference

1. Ed Noll, W3FQJ, "Solar Power," *ham radio*, November, 1974, page 52.

ham radio

simple audio-frequency keyer for RTTY

This easily-duplicated afsk unit produces a stable, smoothly-switched, sine-wave output from a minimum of components

W.H. King, W2LTJ, 5 Midwood Drive, Florham Park, New Jersey 07932

With a minimum of parts and a bit of electronic serendipity, I designed and constructed the audio-frequency-shift-keyer described herein. I received excellent and gratifying reports as well as queries on its circuitry from experts on the air and friends who have used my "spare" on both afsk a-m on two meters and fsk via ssb-afsk on the low bands. Most don't believe me when I tell them that it uses only one transistor and no filter or other complication. The device, however, doesn't know that, so it continues to function and to enjoy the flattery of its more complicated counterparts. Consult the circuit diagram (fig. 1) and you'll see immediately that it is a gutless thing, with only a few parts.* I almost forgot to mention that it also puts out a pure sine wave!

circuit description

The power supply is derived from the loop current (*typically* 60 mA), and when the loop current disappears during spacing, the electrolytic supplies the 4 mA required by the oscillator. Since the device is insensitive to voltage changes, the small drop in supply voltage is not detectable.

As will be developed in the data tables to follow, power supply voltage changes from 15 to 30 volts have less than 6 Hz effect on the frequency and only a 5 per cent effect on the audio frequency output voltage. The oscillator is a Hartley type, using a resistance loaded inductor with a Q of ten. The "grid-leak" has no capacitor across it. These points, in addition to the current regulating nature of the jfet, are what allows transient-free frequency-shift keying.

The opto-electric coupler keys a diode-switched capacitor to change from mark to space frequencies. The coupler is required for isolation and to obtain the very high impedance needed for good switching.

*A printed-circuit board and the components for the afsk can be obtained from Varco Devices, Drawer 8, Stirling, New Jersey 07980.

Note that you could get away with one logic diode in lieu of the diode bridge if you connect the RTTY loop up with the proper polarity. The bridge arrangement is better, as anybody can wire it up and get the correct polarity the 2975-Hz space frequency by unwinding turns from the toroid.

You can tune the unit on the RTTY loop, or you can use an equivalent source of power of 18 volts, current limited with a resistor to 60 mA. What-

table 1. Audio frequency and rms voltage at top of toroid as a function of supply voltage. Above 15 volts both frequency and audio output are independent of supply voltage.

	850 Hz shift tones				170 Hz shift tones			
	marking		spacing		marking		spacing	
power supply volts	freq Hz	audio volts	freq Hz	audio volts	freq Hz	audio volts	freq Hz	audio volts
10.0	2100	13.3	2900	13.5	2096	13.3	2264	13.3
15.0	2125	18.5	2966	18.7	2121	18.7	2294	18.5
20.0	2134	19.7	2983	20.5	2128	20.1	2302	20.1
25.0	2133	20.1	2986	20.3	2128	20.3	2303	20.0
30.0	2131	19.6	2987	19.6	2130	20.1	2304	19.3

on the circuit, thus providing protection at the same time!

It is no surprise that the 88-mH toroid and the capacitors aren't "on the button" items; thus, you will have to tune the circuit to the appropriate frequencies by trial and error. For 850-Hz shift keying between 2125-Hz mark and 2975-Hz space, the values for C1 and C2 are 0.0317 μF and 0.0330 μF, respectively. For 170-Hz shift the values are 0.0555 μF and 0.0092 μF. Please note that the toroid in this case had an inductance of 86.7 mH. You could never be so lucky as to get an identical one, but you can come close if you start with a capacitor of about 0.033 μF and tune to ever you use, I suggest that you connect a temporary short across the LED in the MOC1002 (pins 1 and 2). This will put the unit in the spacing condition and you can now tune C1. Next, remove the short and tune C2 to the mark frequency. Since there may be a little interaction, recheck it, and when you are satisfied solder the capacitors in place. If you desire both shifts, you can add a dpdt switch and the two additional capacitors.

The only other adjustment depends upon your audio output requirements. Each turn of output secondary you wind on the toroid gives about 20 mV of audio output for your microphone

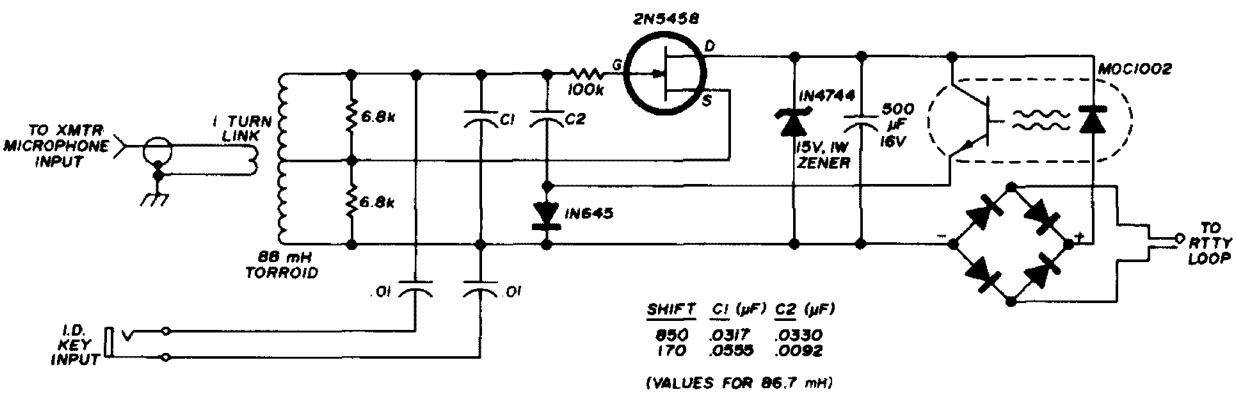


fig. 1. Schematic diagram of the afsk Keyer. Values for C1 and C2 are approximate; oscillator must be tuned to frequency by removal of turns from the 88 mH toroid.

table 2. Audio frequency as a function of shift, loop current and supply voltage.

power supply volts	spacing zaro mA 850	spacing zaro mA 170	marking 50 mA 850	marking 50 mA 170	marking 70 mA 850	marking 70 mA 170
5.0	2784	2209	2503	2053	2050	2047
7.5	2891	2260	2093	2090	2094	2090
10.0	2934	2280	2112	2107	2111	2109
12.5	2957	2291	2130	2118	2120	2117
15.0	2967	2296	2140	2122	2125	2122
17.5	2974	2298	2140	2124	2127	2124
20.0	2975	2299	2139	2124	2128	2125
22.5	2976	2300	2138	2125	2128	2124
25.0	2977	2300	2136	2125	2128	2125
27.5	2978	2300	2135	2125	2127	2125
30.0	2978	2300	2133	2125	2128	2126
ideal	2975	2295	2125	2125	2125	2125

Note: Frequencies at 60 mA are within a few Hz of those shown at 70 mA.

input. In my case one turn was enough, so you can be accordingly cautious.

performance

I don't intend to defend my design with great amounts of data which may prove difficult for you to verify. But let me say this: on 170-Hz shift you can't find any defect or transient problem as determined by scope, ear, or RTTY expert; on 850-Hz shift there is a bit.

Recently, I borrowed a Western Electric Telegraph Transmission Measuring Set (model 164C2) and used it to check the above conclusions. I played a test tape into the TD to key the loop and thus the afsk. The tones were simultaneously demodulated with a terminal unit which keyed a second loop containing the distortion meter and a printer. The 170-Hz shift was — as advertised — perfect, and the 850-Hz shift mode showed about one percent distortion that could be blamed on the afsk.

The data in tables 1 and 2 quantify my statements about the frequency stability and constant-amplitude audio output of the circuit. In order to make these tests, the circuit was altered by applying static voltages, then measuring the resulting frequencies and voltages.

ham radio

the ham notebook

TTL crystal oscillators

Since building a frequency counter, I've been intrigued by TTL crystal oscillators because of their simplicity. Previously I had not been able to obtain reliable performance at frequencies above a few MHz, but I recently ran across a circuit that purrs like a kitten at 14.833 MHz, the highest frequency crystal in my junk box. While checking out this circuit, the possibility of frequency modulating a TTL oscillator came to mind. The circuit is shown in **fig. 1**.

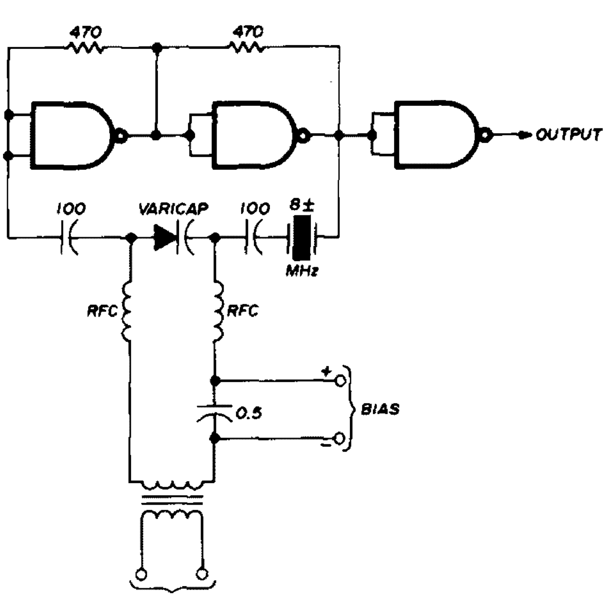


fig. 1. Frequency-modulated crystal oscillator uses TTL gates and varicap diode. Performance curves are shown in fig. 2.

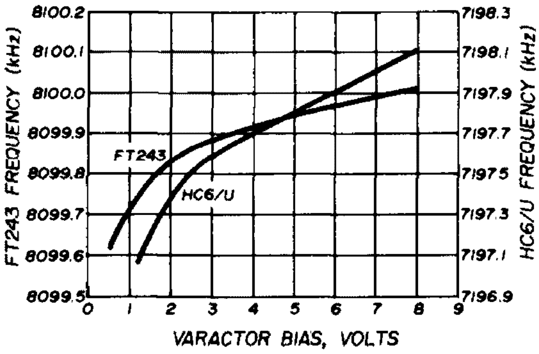
The varicap I used was one of Radio Shack's specials, since that was what was in the junk box. With no audio, the bias was varied to obtain voltage/deviation curves with two available crystals, an FT243 at 8.1 MHz and an HC6/U at about 7.2 MHz. The two curves are shown in **fig. 2**. A bias of 4 volts was chosen.

Since I hadn't been bitten by the 2-meter fm bug at the time I first built this circuit, there was nothing in the house to monitor the output except a six-band general-coverage receiver which covered the 2-meter band along with police and other frequencies. The 18th harmonic of the oscillator was strong and the audio quality seemed to be surprisingly good. I had KØPHF's Heath Twoer on the bench, and with one connection change (and a different crystal) the output of the oscillator was plugged into the Twoer's crystal socket. It drove the Twoer nicely, and to keep it in the sphere of amateur radio, a contact was made through the local 34/94 repeater.

The Twoer first stage, a tripler, seemed to flourish on the harmonic content of the TTL oscillator. Succeeding stages should do a fair job of curtailing final harmonic output. However, this must be carefully checked before any long-term 2-meter operation as additional filtering may be required.

George Chynoweth, WØJVA

fig. 2. Deviation of frequency-modulated TTL oscillator shown in fig. 1.



filter alignment

Moore covers a lot of territory in his recent article[1], but he couldn't include everything. However, one point does need to be expanded: using the spectrum analyzer and noise generator for matching and aligning filters. I have been using this method for several years.

I first used the noise generator in my Omega-T antenna bridge in lieu of a good sweep generator that would cover the i-f frequencies. A Heath SB620 Scanalyzer was connected behind or ahead of the filters under test. Unfortunately, the SB620 covers only the i-f it is built for. I was interested in an i-f of 5645 kHz and the SB620 was equipped with those coils (various frequency coils are supplied in the SB620 kit).

The object of all this was to properly match and align a Drake R4B receiver after installing an 8-pole, 3.7 kHz filter in place of the original, rather wide, 4-pole, 8 kHz filter. More recently the 8 kHz filter was removed from an R4C and replaced with a 5 kHz filter.

To align the filter I coupled the noise bridge into the antenna terminal of the receiver, and the AVC was turned off. The SB620 is coupled very loosely to the grid of the first tube after the filter which is being adjusted. Observe the lower corners of the bandpass curve. You will probably see something like fig. 3, curve A. But what you want, and can get by adjusting the coupling components, is curve B.

What you may not suspect is that the SB620 can be placed ahead of the filter to display the "suck-out" of the filter (its low-impedance swamping of the incoming noise). This curve, fig. 4, is just about the reciprocal of fig. 3. The area between the solid and broken lines is what you have gained by proper matching and isolation.

The curve in fig. 4 also illustrates what happens when the SB620 is con-

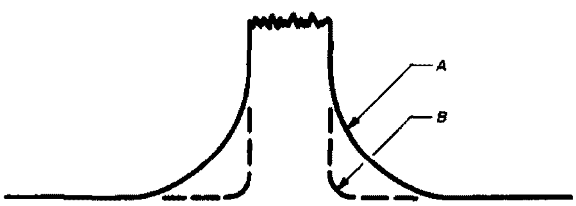


fig. 3. Display of filter response with noise generator input and SB620 installed after the filter. Curve B can be obtained with careful matching and tuning.

nected as a pan adapter in front of a good filter which has a fairly low impedance. This results in off-frequency signals appearing stronger than those within the passband of the filter. This is rather disconcerting when you wish to check the frequency to which you are tuned: it will look like a clear spot and a good place for a CQ when, in fact, there may be signals there.

Arthur E. Lux, W7UC

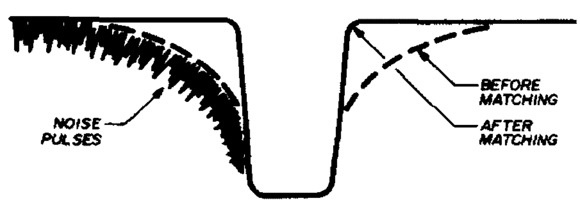
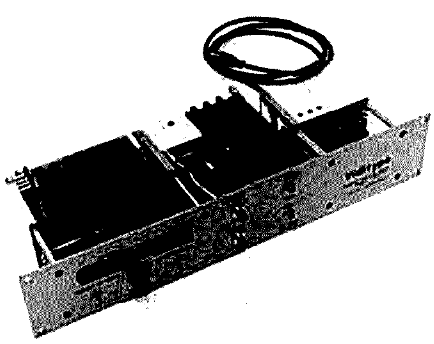


fig. 4. Display of filter "suck-out" with SB620 installed in front of the filter. This curve is nearly a reciprocal of that shown in fig. 3.

1. Ray Moore, "Designing Communications Receivers for Good Strong-Signal Performance," *ham radio*, February, 1973, page 6.

new products

RTTY video display unit



Leland Associates has announced a new RTTY video display unit, the model 872 Viditype, which will provide a video display of RTTY signals on *any* standard television receiver's unused channel. No modifications are required to the TV receiver – simply connect the unit to the antenna terminals. The model 872 is directly compatible with all common RTTY terminal units and features CR-LF on line feed signal, automatic CR-LF on 40th character or a space in positions 36 to 40, automatic page unshift at the end of the 25th line, manual clear (starts print at upper left screen), selectable down-shift on space, manual letters key, and print suppression on all nonprinting functions. Character format is 40 characters per line, 25 lines of display and 1000 character display capacity. The video system uses a crystal-controlled sync generator with a video bandwidth of 3 MHz and 30 Hz refresh rate. Output is 2000 microvolts at 50 ohms on TV channels 3 to 6.

The model 872 video display unit is priced at $450 in kit form ($550 assembled and tested) plus $3.00 shipping and insurance. Specify Baudot or ASCII input. For more information, write to Leland Associates, 18704 Glastonbury Road, Detroit, Michigan, 48219, or use *check-off* on page 94.

triad catalog

Triad-Utrad's new *1975-76 Replacement Catalog and Television Guide for Transformers* is now available. The 70-page catalog features several hundred replacement transformers including color TV components, deflection yokes, flybacks, vertical outputs and filter chokes, as well as power, filament and audio transformers. Copies of the catalog are available on request from Steve Fisher, General Manager, Triad Utrad Distributor Services, 305 North Briant Street, Huntington, Indiana 46750, or use *check-off* on page 94.

450-MHz fm transmitter and power amplifier

VHF Engineering has recently announced the availability of a new 450-MHz fm transmitter and companion 10-watt, 450-MHz power amplifier. For the first time, simple kits are available to permit fm operators to get on 450 MHz without relying on expensive new or surplus equipment. Previously, the uhf fmer had to purchase surplus tube-type uhf fm equipment which was expensive (and difficult to maintain), or

More Details? CHECK–OFF Page 94

he had to purchase new fm gear designed for the amateur market. The new gear is much more reliable than the older surplus, but it is very expensive.

VHF Engineering has announced the availability of a simple 1-watt, 450-MHz transmitter kit and a 10-watt 450-MHz amplifier kit designed for construction by the average amateur. These kits are relatively easy to build and do not require sophisticated test equipment. The kits are fully solid state and use readily available components. An experimenter who builds these kits will be able to maintain them himself, a distinct advantage over purchasing a wired and tested unit.

The 450-MHz transmitter consists of five simple stages starting with a varactor-modulated crystal oscillator using crystals in the 18-MHz range. The oscillator quadruples to 55 MHz and drives the first of three doublers. The first two doublers use 2N3866 transistors in a standard doubling configuration. The last doubler uses a 2N3553 and delivers output to the final on 450 MHz. The final amplifier transistor is a 2N5913 operating straight through at 450 MHz, delivering 1 watt output out on the 450 MHz fm band.

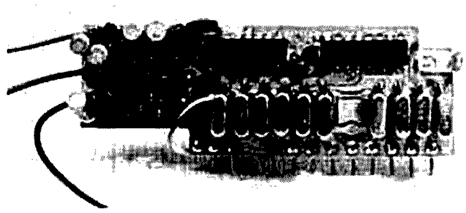
While one watt may be sufficient for some applications, additional power really helps in rough terrain when using a repeater and is an absolute must when operating direct on 450 MHz. The VHF Engineering 450-MHz, 10-watt amplifier will supply this extra power at nominal cost. The power amplifier is well designed and very easy to build. Most experimenters should be able to complete it in less than two evenings. The power amplifier uses two balanced-emitter uhf transistors and delivers slightly more than 10 dB gain. For one watt input, the minimum output is 10 watts.

Each VHF Engineering kit consists of top quality components and epoxy-glass circuit boards. The instructions are

straight forward and include sections on theory, construction, operation, and trouble-shooting. Also available from VHF Engineering are a companion solid-state 450 MHz receiver and a solid-state 450 MHz repeater.

The VHF Engineering 450-MHz transmitter kit sells for $39.95, and the 10-watt amplifier sells for $39.95. VHF Engineering kits are available from dealers or direct from the manufacturer at 320 Water Street, Binghamton, New York 13902. For more information, use *check-off* on page 94.

fm scanner



Topeka FM Communications has just released a ten-channel scanner designed for use with Regency's HR-2 series radios. This unit will also work on Regency marine radios MT-15, MT-25 and Aquaphone, and is designed to rapidly scan ten channels and lock on any frequency which has a strong enough signal to open the receiver's squelch circuit. A scan-lock feature prevents continued scan due to a momentary signal loss.

The scanner features a priority channel which is selected by the channel selector switch. While receiving on any frequency, the scanner periodically checks the priority channel and returns to it if a signal is present. This feature is ideal for the receivers that must monitor emergency frequencies. Delayed scan after transmit allows time for an answer before the scan is resumed. A simple modification allows selective channel bypass. The scanner is priced at $52.50. Order from Topeka FM Communications, Inc., 125 Jackson, Topeka, Kansas, 66603.

More Details? CHECK—OFF Page 94

$1.00

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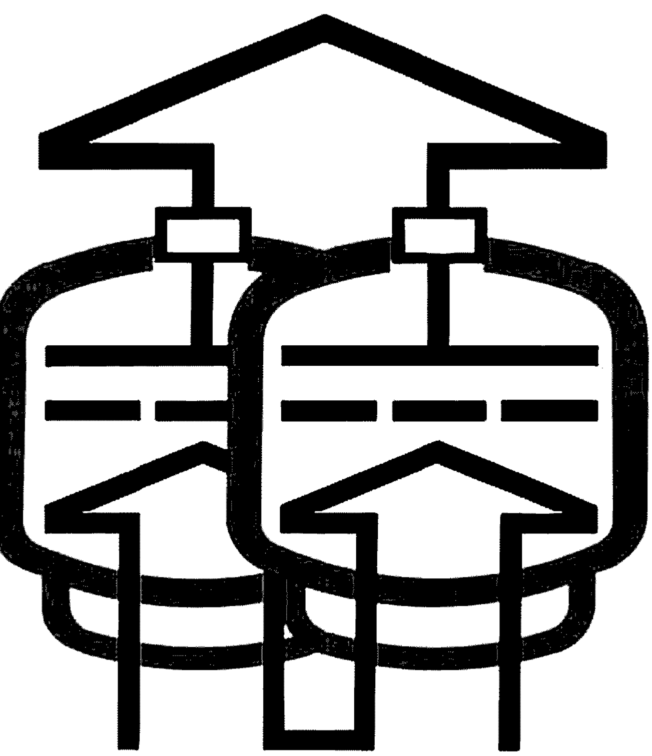
ham radio magazine

SEPTEMBER 1975



inductively-tuned six-meter kilowatt

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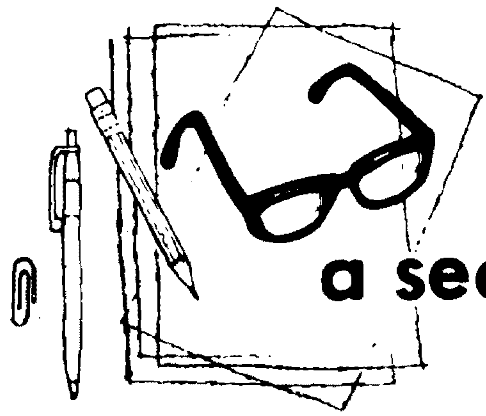


a second look

by jim fisk

Recently there has been rising concern over the possible harmful effects to living tissue due to heating by electromagnetic radiation in the frequency range from 10 MHz to 100 GHz. Although microwave engineers have been aware of the potential hazards of working around high-power microwave transmitters for 25 years or more, few people expressed much concern about the possible radiation hazards of lower-frequency equipment. However, from 150 to 1200 MHz the internal body organs are susceptible to damage from rf heating, and the eye is especially prone to damage from radiation above 1000 MHz.

The scientific community is not at all satisfied that there has been sufficient research for formulating rf exposure standards, but based on present knowledge, various governmental and industrial organizations involved in establishing radiation safety standards have recommended exposure limits referred to as Radiation Protection Guide Numbers (RPGN). At the present time the generally accepted RPGN value is 10 milliwatts per square centimeter, and the Occupational Safety and Health Administration (OSHA) has promulgated a standard which limits exposure to power densities greater than 10 mW/cm^2 of body area.

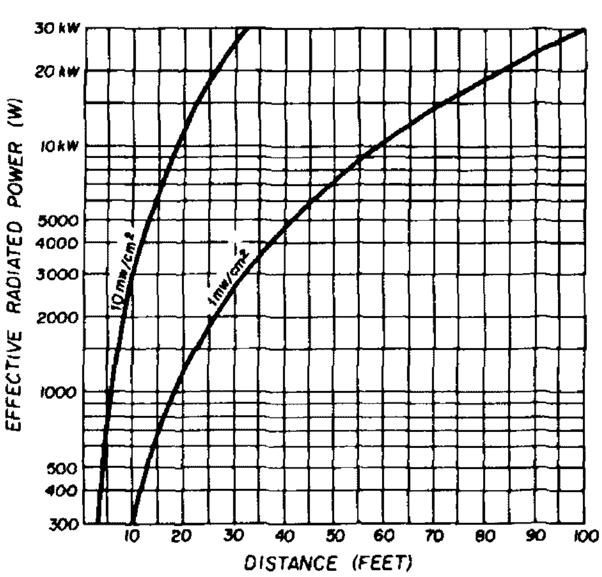


fig. 1. Minimum distances from radiators at which Radiation Protection Guide Numbers are not exceeded vs effective isotropic radiated power (EIRP).

Authorities generally agree that rf power levels one-tenth the OSHA standard (1 mW/cm^2) do not have any noticeable effect. Using this as a basis, what are typical safe distances from a high-frequency amateur antenna?

Since the safe distance from an antenna depends on its radiated power in a given direction, the most direct approach to finding the distance is to use a graph such as that shown in **fig. 1**. These curves are based on an isotropic radiator so antenna gain (power ratio, *not* dBi) must be factored in for the practical case. A half-wave dipole, for example, has a power gain of 1.64 over an isotrope (2.14 dBi). Assuming 1000 watts into the antenna, what is the minimum safe distance? The effective isotropic radiated power (EIRP) is $1.64 \cdot 1000$ or 1640 watts and the distance for a power density of 1 mW/cm^2 is about 24 feet.

Since most dipoles are installed at least 25 feet above ground, they obviously pose little threat at amateur power levels, but what about multi-element beams and stacked arrays? Assuming the array is at the top of a 54-foot tower, and not facing into a building, an EIRP of about 29 kW would be required to produce 1 mW/cm^2 in the center of the main lobe 100 feet away. With 1000 watts into the antenna, this corresponds to a power gain of 29 or 14.6 dBi.

Few amateur antennas exhibit this much gain, but EME operators who use large arrays or big parabolic reflectors should use caution. A 30-foot dish with only 10 watts input at 432 MHz, for example, is hazardous at distances of less than 18 feet!

Jim Fisk, W1DTY
editor-in-chief

presstop

AN IMPORTANT STORY is still Prose Walker's decision to step down as chief of the Amateur and Citizens Division on July 31st. Deputy Division Chief Dick Everett is presently serving as Acting Division Chief.

In The Long Term the question of who'll be filling Prose's shoes is a tough one. Ideally the job requires an engineer-ham with strong managerial experience — plus appropriate Civil Service status. Though the new chief is likely to come from within the Commission it isn't an absolute must. However, candidates without Civil Service Ratings must establish their qualifications with Civil Service before they can be considered for the post.

CB RESTRUCTURING — DOCKET 20120 — has been partially decided in a recently released Report and Order. Most important change is a sharp relaxation of prohibited communications effectively permitting the hobby use of CB that has already characterized the service legal or not.

Also Relaxed are identification procedures, with an operator required to give only his callsign instead of both; the "quiet" period between five-minute series of transmissions has been reduced from five minutes to one minute; inter-licensee communications are now permitted on all 23 channels. Channel 9 continues to be classified as the emergency calling channel, and channel 11 has been designated a general calling channel.

Notable Emissions from this initial Report and Order on the Docket are the expansion to additional channels above the present band and the proposed conversion to all SSB. Though action on these can be expected eventually, it will probably rest to some degree on what happens with Class-E CB.

MANDATORY REVIEW OF LOGGING TAPES for repeaters operating under "automatic control" will not be required as a result of recent FCC decision. Logging tapes must still be made and kept for 30 days, and if any operating problems are reported to the control operator or licensee during that period the tapes must be reviewed and corrective action taken within 72 hours or the system must shut down.

This Easing Of Requirements for repeater operators takes most of the sting out of the Commission's Report and Order — mandatory review had been the principle objection raised by the repeater fraternity.

AMSAT MEMBERS not planning to attend AMSAT's annual meeting at the ARRL Convention in September are reminded that their ballots for the Director's election are needed to insure a quorum. The annual meeting is scheduled Sunday, September 14, at Reston, Virginia.

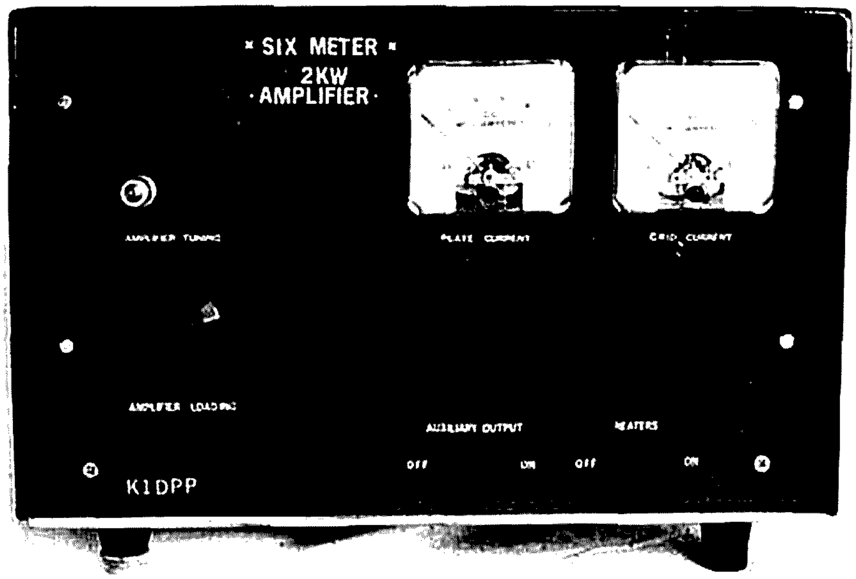
New Two-Color Bumper Sticker Decals are now available from AMSAT. They are being sent free to new members and renewing old members, or sold three for $1 plus SASE from AMSAT, Box 27, Washington, DC 20044.

TV TUNER WITH MEMORY has some interesting design features for Amateur radio applications. Announced in July by General Instrument, the tuner uses a non-volatile memory chip that contains 100 lines of 14 bits each. Each line can be programmed for one TV channel, and when needed the memory information is fed to a 14-bit CMOS D/A converter to produce an analog voltage which is applied to a varactor diode and tunes in the station.

HAM GEAR SHORTAGES have been plaguing dealers for some time and are likely to get worse before they get better. The major cause of the problem is the CB explosion, since many manufacturers — particularly those in the Far East — supply both the Amateur and CB markets and it pays them to put their major effort in the market with the most money.

TYPE ACCEPTANCE OF AMATEUR GEAR is still a hot issue around the FCC. Don't be surprised to see a Notice of Proposed Rule Making come out of the Commission soon.

Continued Abuses By CBers using Amateur transceivers and manufacturers building "broadband" linears for the "Amateur Radio market" that just happen to deliver full output with only 4 watts drive (on ten meters, of course) have pretty well forced the FCC to act.



inductively-tuned six-meter kilowatt

Construction details for a high performance grounded-grid six-meter linear using 3-500Zs

Donald Cook, K1DPP, 628 Rindge Road, Fitchburg, Massachusetts

At one time or another most serious six-meter operators have appreciated the value of having a high-power amplifier for extending their communications range – scatter, extended ground wave and aurora contacts are enhanced considerably by the proper use of an amplifier. I decided to build a two-kilowatt PEP linear amplifier around a pair of grounded-grid Eimac 3-500Zs because I have a Swan 250C which easily provides the 100 watt PEP input drive requirement. In addition, I have two other rf power amplifiers using 3-500Zs which have given me trouble-free service for the past several years. A spare pair of tubes that were sitting on the shelf helped convince me to go in this direction.

After acquiring operating manuals from different manufacturers and going through the various amateur magazines, scanning articles and circuits which used 3-500Zs, I built the six-meter kilowatt described here.

The 3-500Z tubes are air-cooled power triodes designed for zero bias operation and are rated to 110 MHz. They are cathode driven with the grids at rf and dc ground. The tuned-cathode input circuit (**fig. 1**) provides good linearity and minimizes the drive requirements. The filaments are isolated from rf ground with the high-current bifilar rf choke, RFC1. The tuned plate circuit in this amplifier, however, is a bit unusual – the normal pi-network tuning capacitor has been replaced by an inductively-tuned or shorted-turn plate tank coil. Most circuit losses at 28 MHz and up are due to residual circuit capacitance and much of this undesired effect is due to the pi-network input tuning capacitor commonly used at lower frequencies.[1]

In this circuit the pi-network input capacitance, C5, consists of stray circuit and tube capacitance.

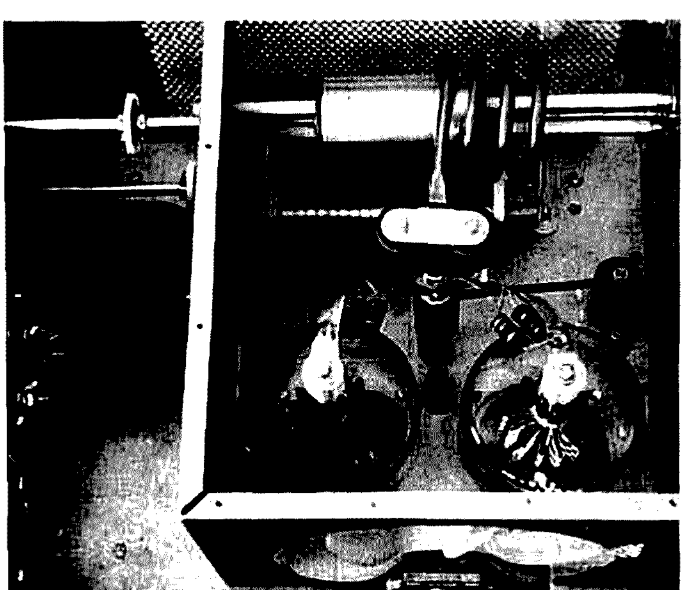
The control relay, K1, which switches the amplifier into the circuit and shorts out the 10k, 10 watt resistor, allowing normal plate and grid current to flow, is operated from a low current dc supply derived from a winding on the filament transformer. External interlocks (not shown here) make it impossible to apply plate voltage without first turning on the filaments and blower.

The two meters on the front panel of the amplifier shown in the photographs are used to read the grid and plate circuits. Since the meters I used had 100-milliampere movements, I used resistance shunts to provide a multiplier of five for the grid current meter (500 mA) and ten for the plate current meter (1 ampere).

construction

The pi-network inductor is wound from ¼-inch (6.5mm) diameter copper tubing and the shorted turn is made from a section of 1½-inch (38mm) copper water pipe. This slug is moved in

Plate compartment of the six-meter kilowatt showing construction of the pi-network inductor, parasitic chokes and RFC2. The neutralizing capacitor is located between the two tubes. The pi-network loading capacitor is positioned underneath the tunable inductor.



and out of the inductor to tune the plate tank circuit to resonance. Construction details for this assembly are shown in fig. 2.* Although I used a simple spinner knob on the inductor tuning screw, a turns-counting dial could be used to provide a logging reference for various operating frequencies.

The grid pins on the 3-500Z sockets are directly grounded to the chassis with ¼-inch (6.5mm) wide copper strap. One end of the strap is soldered to the socket pins and the other end is attached to the chassis with 6-32 (M-3.5) screws and nuts. Make sure that both contact surfaces are clean for low-resistance, trouble-free grounding. The homebrew filament choke consists of 12½ bifilar turns number-12 (2.1mm) Formvar on a ½-inch (13mm) diameter ferrite rod (suitable filament choke kits are available from Amidon Associates†).

The shunt-feed rf choke, RFC2, is connected directly across the plate tank circuit, so it must provide good isolation over the entire six-meter band. Hand winding, as described here, is highly recommended as no commercially available rf choke is apt to provide as good performance. The winding is most easily accomplished by feeding *two* number-20 (0.8mm) wires through one hole in the coil form and winding a bifilar coil of 30 turns, ending at the other hole in the form. Remove *one* of the bifilar windings and you will have a tight, evenly spacewound winding that makes an excellent six-meter choke. The cold end of the rf choke is bypassed to ground with a 500 pF TV-type doorknob capacitor.

The 500-pF dc blocking capacitors, C3 and C4, are mounted between two

*Major components for the shorted-turn inductor are available from Edward A. Stoltzfus, Engineering Machinist, Beacon Light Road, Parkesburg, Pennsylvania 19365.

†Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607. Filament choke kit is $2.50, postpaid.

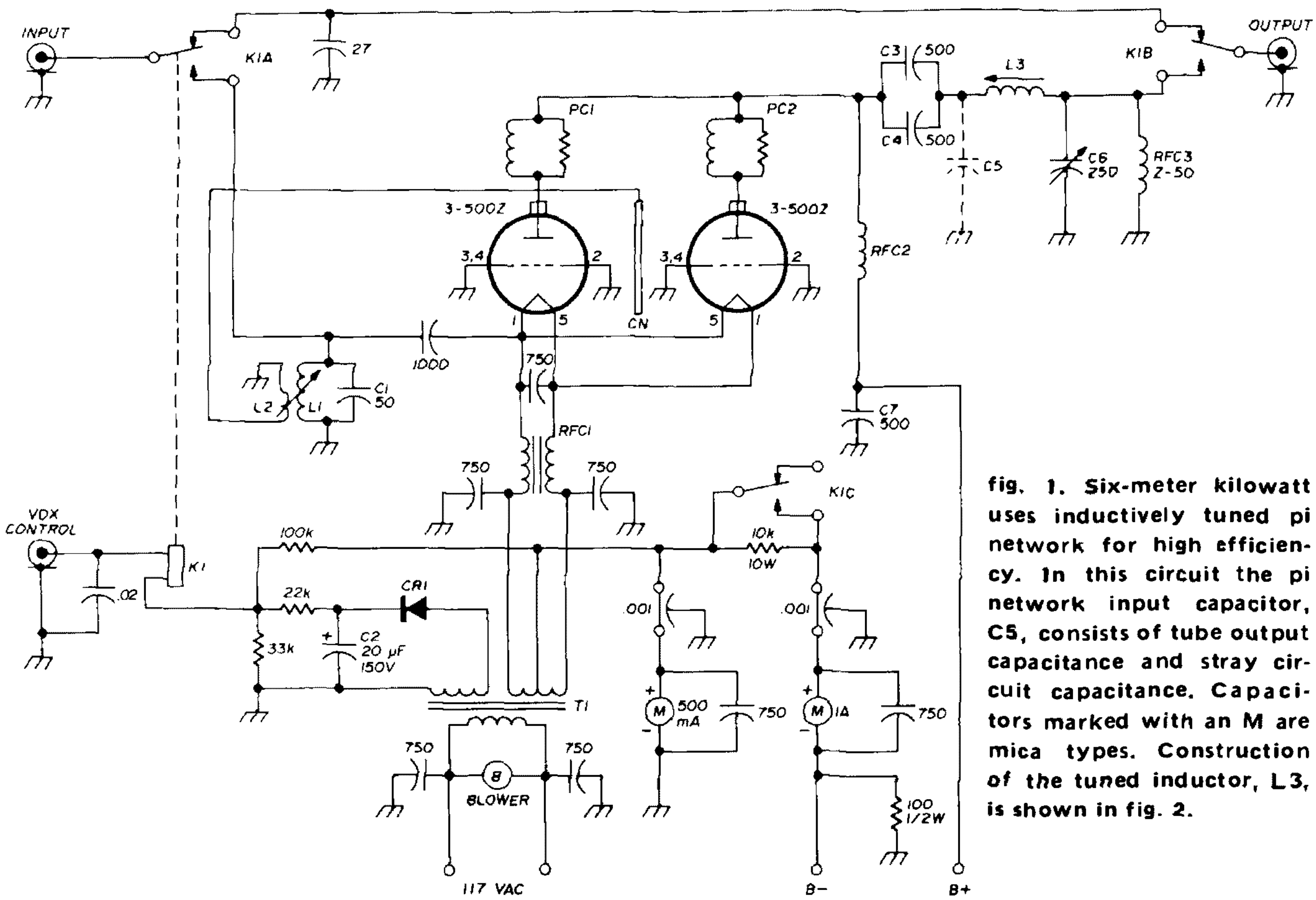


fig. 1. Six-meter kilowatt uses inductively tuned pi network for high efficiency. In this circuit the pi network input capacitor, C5, consists of tube output capacitance and stray circuit capacitance. Capacitors marked with an M are mica types. Construction of the tuned inductor, L3, is shown in fig. 2.

C1	50 pF compression mica trimmer (ARCO 462). Adjust L1 for resonance with C1 compressed 90 percent
C3,C4	500 pF, 5 kV doorknob capacitor (Centralab 858S)
C5	residual circuit and tube output capacitance
C6	250 pF air variable (E. F. Johnson 154-1)
C7	500 pF, 10 kV doorknob capacitor (TV type)
CN	Neutralizing capacitor, copper strap ½" (13mm) wide by 3½" (89mm) long
CR1	Silicon diode, 600 PIV, 1 A
K1	110 Vdc relay, 3 pole, double throw (Potter & Brumfield KA14DG or Heath 69-55
L1	3 turns no. 14 (1.6mm), airwound, ½" (13mm) diameter, 5/8" (16mm) long
L2	2 turns no. 18 (1.0mm) insulated wire, ½" (13mm) diameter, placed between turns of L1 (note polarity)
L3	3½ turns ¼" (6.5mm) copper tubing, 3" (76mm) long, 2-1/8" (54mm) inside diameter. Tuning slug details are shown in fig. 2.
PC1,2	Parasitic suppressors, Each consists of three 50 ohm, 2 watt resistors shunted across ½" (13mm) wide copper plate strap. Install as close to the plate cap as possible (see photo)
RFC1	12½ turns no. 12 (2.1mm) Formvar, bifilar wound on ½" (13mm) diameter, 3½" (89mm) long ferrite rod (Amidon Associates filament choke kit)
RFC2	30 turns no. 20 (0.8mm) enamelled wire, space wound on 3/4" (19mm) round, 3-3/4" (95mm) long ceramic or Teflon rod. Drill holes for wire ends ½" (13mm) and 2-3/4" (70mm) from top or rod
T1	dual secondary transformer. 5 Vac at 30 amps (filaments), 120 Vac (relay K1 dc supply)

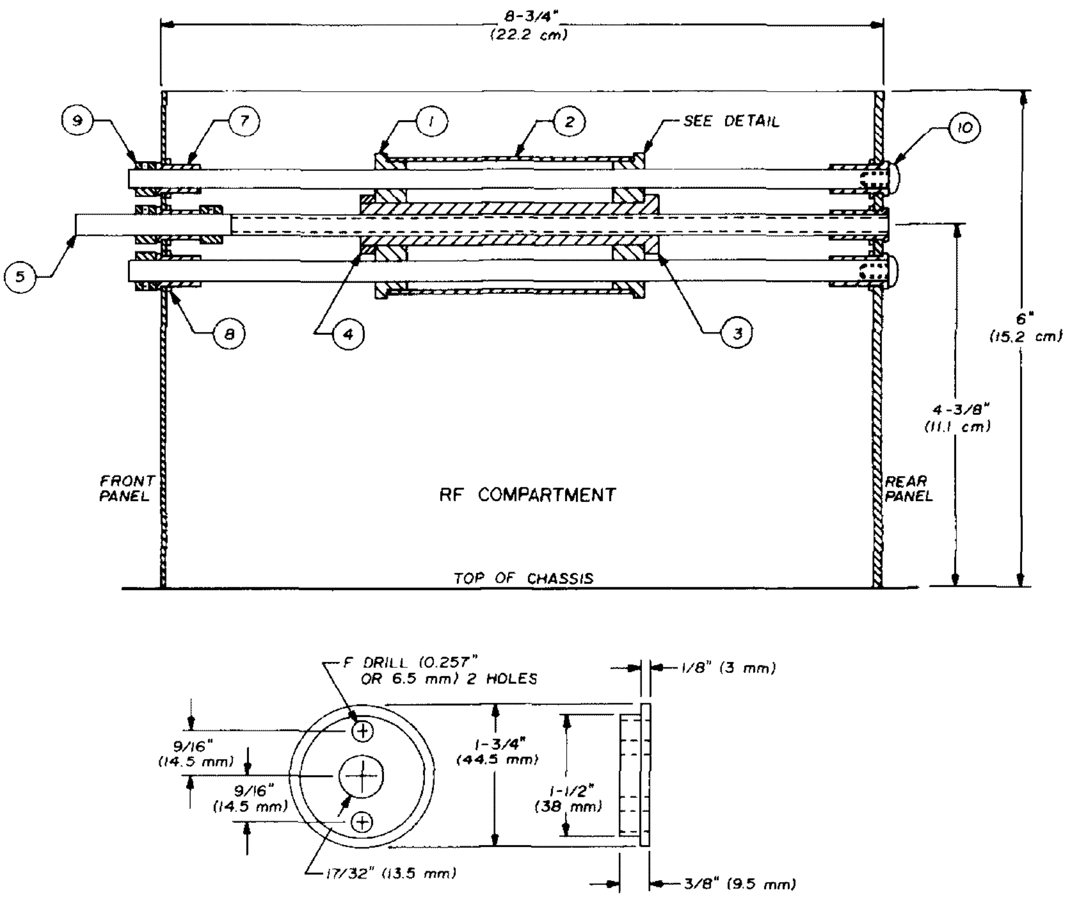
brass plates, one of which is attached to the top of RFC2. The other plate is connected to the pi-network inductor L3 (see photo of the rf compartment).

The amplifier is built on a homemade chassis made from 3/32-inch (2.5mm) thick aluminum sheet and is 1-3/4 inch (4.5cm) high, 12 inches (30.5cm) wide and 13 inches (33cm) deep. The front and rear panels are 1/8-inch (3mm)

thick aluminum, 8 inches (20.5cm) high by 13 inches (33cm) wide. The inner right- and left-hand panels are 1/16-inch (1.5mm) thick aluminum, 7½ inches (19cm) high by 13 inches (33cm) long. These two panels can be made from perforated stock or solid sheets can be drilled to provide for ample air intake and exhaust to cool the plate and filament seals.

The rf compartment is 10 inches (24.5cm) wide, 8-3/4 inches (22cm) deep and 6 inches (15cm) high. The top cover for this compartment is made from perforated aluminum sheet and should be installed at all times to protect the operator from the lethal voltages which are used to operate this equipment.

The bottom cover is 1/16-inch (1.5mm) thick aluminum, 13 inches



1. Insulating disc, 1-3/4" (44.5mm) diameter, 3/8" (9.5mm) long (2 required).
2. Tuning slug. 1-5/8" (41.5mm) OD, 1/16" (1.5mm) wall copper water pipe, 3" (76mm) long.
3. Tuning mechanism, make from brass rod, ½" (13mm) OD, 3-5/8" (92mm) long. Drill and tap for ¼-20 (M7) thread for full length. Thread outside diameter ½-13 (M12) 1" (25mm) on both ends.
4. Brass nuts, ½-13 (M12), split into two pieces about 3/16" (5mm) long.
5. Lead screw. ¼" (6.5mm) diameter brass rod, 9-7/8" (25cm) long, threaded ¼-20 (M7) for 8" (21cm).
6. Guide rods. Plastic or fiberglass rods, ¼" (6.5mm) diameter, 9¼" (23.5mm) long (2 required). Drill and tap one end ½" (13mm) deep for 6-32 (M3.5) machine screw.
7. Panel bushings for ¼" (6.5mm) shaft (6 required).
8. Panel bushing nuts (6 required).
9. Shaft stops, for ¼" (6.5mm) shafts (4 required).
10. Machine screws, 6-32 x ¼" long (M3.5x6.5) (2 required).

fig. 2. Construction of the tuned pi-network inductor, L3. This arrangement requires a minimum of machine work and uses readily available materials.

(33cm) wide by 14 inches (35.5cm) deep with ½-inch (13mm) lips formed on two sides to provide a mating surface for the top cover. The U-shaped top cover is also made from perforated aluminum. The trim around the raw edges on the top and bottom covers is car door protective edging that is available from most department and automotive stores; it is very easy to work and cut with simple hand tools. The vox connector and power plugs are all brought out on the rear panel along with the input and output coaxial connectors.

Cooling air must be provided to maintain the plate seals of the 3-500Zs below 225°C and the filament seals below 200°C. Many 3-500Z power amplifiers are designed around a system of air-system sockets and chimneys, along

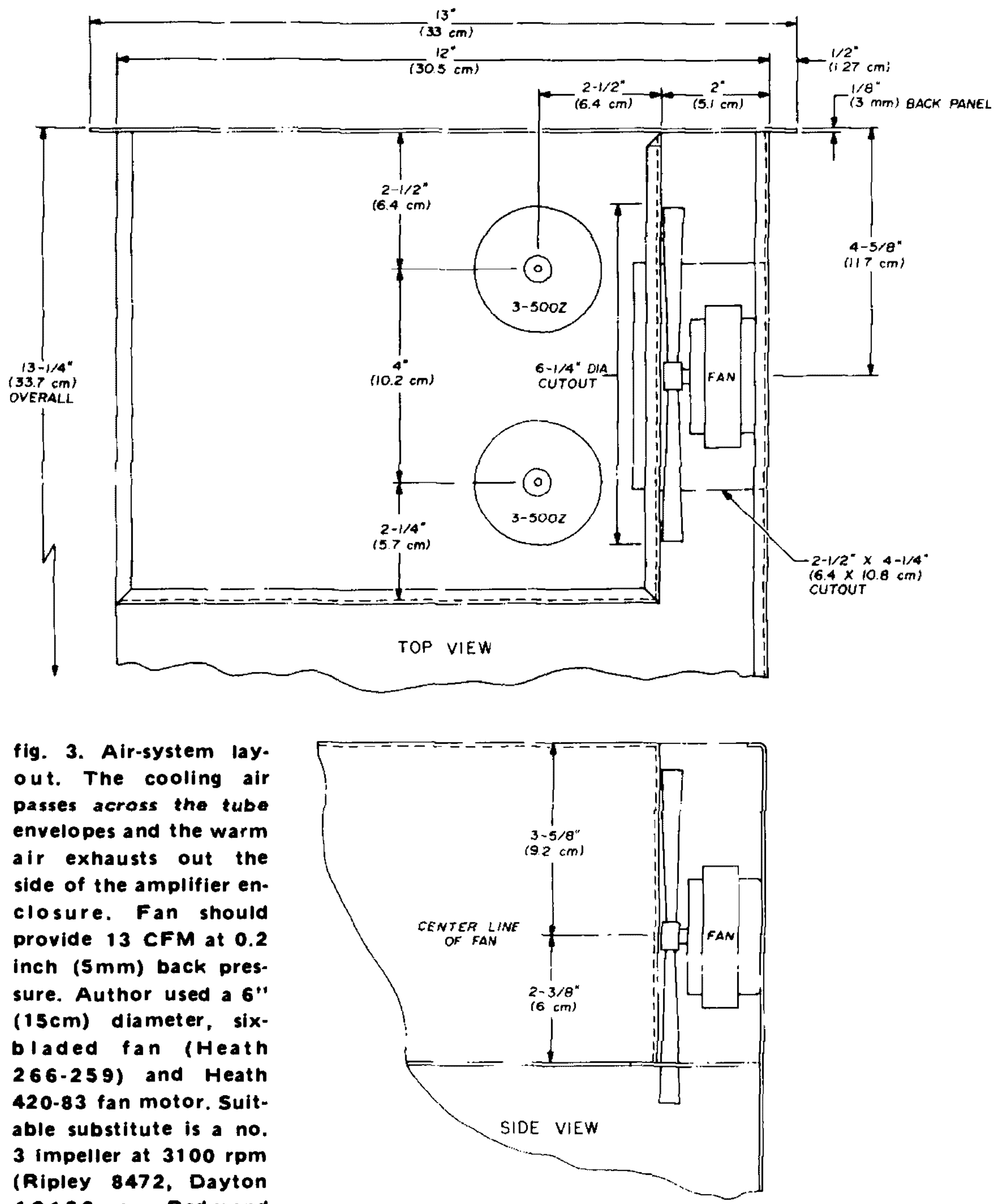


fig. 3. Air-system layout. The cooling air passes *across the tube* envelopes and the warm air exhausts out the side of the amplifier enclosure. Fan should provide 13 CFM at 0.2 inch (5mm) back pressure. Author used a 6" (15cm) diameter, six-bladed fan (Heath 266-259) and Heath 420-83 fan motor. Suitable substitute is a no. 3 impeller at 3100 rpm (Ripley 8472, Dayton 1C180 or Redmond AK-2H-01 Ax).

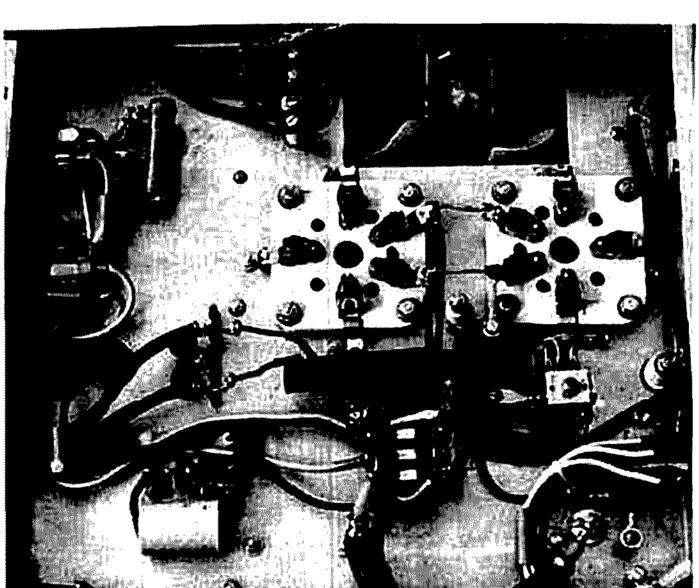
with a centrifugal blower. The noise generated by the blower motor and air movement through the cooling system, however, is very distracting. Extensive tests by Eimac have shown that for CW and ssb operation at legal amateur power limits the 3-500Zs may be adequately cooled by a lateral air blast blown against the tubes by a small, properly positioned rotary fan.[2]

The arrangement of the cooling fan in the six-meter kilowatt is shown in fig. 3. The fan is mounted in the rf compartment wall, between the tubes, in line with the center of the glass envelopes, and blows cooling air across the envelope and plate caps (use a good heat-dissipating plate connector such as the Eimac HR6). The Johnson 122-275-1 ceramic tube sockets are mounted off the chassis about 1/8 inch (3mm) to allow air flow around the base of the tubes to cool the filament pins.

table 1. Typical operating data for the 3-500Z in rf linear amplifier service, class B (two tubes).

Dc plate voltage	1500	2000	2500
Zero signal plate current (mA)	130	190	260
Single tone dc plate current (mA)	800	800	800
Single tone dc grid current (mA)	260	260	240
Two-tone dc plate current (mA)	520	540	560
Two-tone dc grid current (mA)	160	160	140
PEP useful output power (watts)	660	1000	1200
Intermodulation distortion products (dB)	−46	−38	−33

Under-chassis view of the six-meter kilowatt showing placement of major components. The tuned cathode circuit is located at right center.



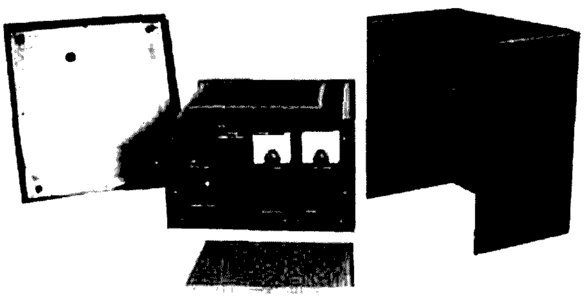
With this arrangement maximum plate dissipation of about 450 watts per tube is achieved for the 3-500Z. While this is about 10 percent short of the maximum rating, dissipation is sufficiently high that the pair of tubes will easily handle the maximum amateur power limit for CW or ssb operation under normal operating conditions. For continuous operation (RTTY or sstv, for example) the power input must be reduced to about 750 watts.

neutralizing

The neutralizing capacitor, CN, is formed from a 0.032-inch (0.8mm) thick copper strip, 1/2-inch (13mm) wide and 3½-inches (89mm) long into an L-shape with the foot of the L about 3/8-inch (10mm) long. The foot is drilled for mounting to the threaded stud of a small ceramic feedthrough bushing. The feedthrough bushing is centrally located between the two 3-500Zs and in line with the socket mounting holes toward the rf choke(see photograph). This capacitor is adjusted to lean in toward the center line of the tubes.

Inductor L2 is inserted between the turns of L1, the input circuit coil, and adjusted until the amplifier is neutralized (stabilization occurs when maximum grid current, maximum output

and minimum plate current all are reached at one setting of the plate tank circuit). With the amplifier turned on but with no drive signal, grid current should be zero. I used an insulated rod to adjust L2 through a small hole in the bottom compartment. The neutralizing capacitor should be adjusted with the power turned off.



Two views of the power amplifier which show the various pieces of metalwork which were used in its construction.

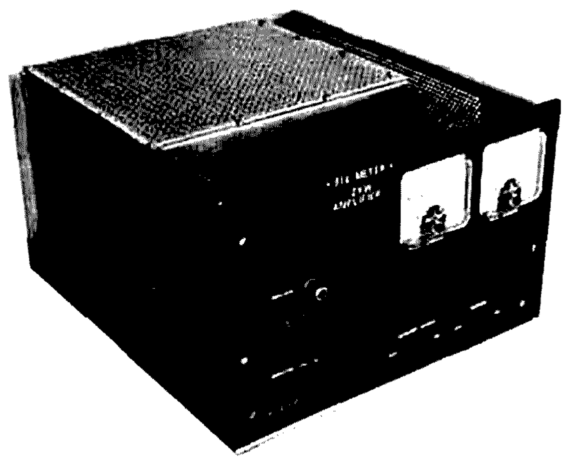
tune up

Before applying any voltages to the amplifier, carefully check all the wiring. Then install the two tubes and use a grid-dip oscillator to approximately set the tuned cathode circuit and pi network output circuit to the desired operating frequency. I used 50-ohm resistors across the input and output coaxial fittings during the grid-dipping operation and made sure that all relay contacts were closed.

After the amplifier has been cold tuned with the grid-dip meter, remove the 50-ohm resistors and connect the six-meter exciter to the input and a dummy load (or antenna) to the output. A monitor scope or relative power meter (an swr bridge works well) will provide a good output indication during initial testing of the amplifier.

First tune up the exciter for full normal CW output with the amplifier switched out of the line. Reduce exciter output, apply power to the amplifier and tune the amplifier for maximum output with reduced B+. Gradually increase the excitation and high voltage to get the feel of a very smoothly tuning six-meter power amplifier.

For CW operation with a 2500-volt power supply tune the plate circuit and adjust the exciter drive for a plate current reading of about 400 mA (about 125 mA grid current). I am using a power supply with a tapped primary which I can switch from 2300 volts for CW and tuning, and then to 2800 volts for ssb operation.



I usually tune for maximum output with the lower voltage and then switch to the higher voltage for ssb. Apply ssb drive from the exciter and advance the microphone gain control for an average plate current reading of 350 mA. An occasional voice peak may boost this up to 600 mA. A monitor scope will show you more than any plate meter can be expected to, and should be used at all times when you are running high power. The manufacturer's operating data for the 3-500Z is published in most handbooks and should be consulted as a guide for proper use of these tubes in amateur service.

references

1. William Orr, W6SAI, "Inductively-Tuned High-Frequency Tank Circuit," *ham radio*, July, 1970, page 6.
2. William Orr, W6SAI, *Radio Handbook*, 19th edition, Editors and Engineers, Indianapolis, 1972, page 22-35.

ham radio

tunable RC notch filter

Discussion of an RC notch filter which can be tuned with a single variable resistor

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Of the host of RC notch filters that have been devised, the twin-T (also called parallel-T) has enjoyed the greatest popularity by far. This is surprising because, in order to make this circuit adjustable or tunable, three components, either resistors or capacitors, must be varied simultaneously. To make matters worse, these three components don't all have the same value as shown in the basic circuit of fig. 1. One of the resistors is one-half the value of the other two, and one capacitor is twice the value of the others. Therefore, an adjustable twin-T notch filter requires a three-gang potentiometer, and the tracking and alignment problems can be troublesome if not expensive.

There is another circuit, however, which has performance comparable to the twin-T, but which can be tuned over a wide frequency range by means of a single potentiometer. Although this circuit has been around for about 20 years, it has seldom appeared in the electronics literature and is not widely used.

To my knowledge, this tunable notch filter first appeared in print in 1955 when it was mentioned by Henry P. Hall of the General Radio Company in the September issue of *IRE Transactions on Circuit Theory*. He discussed it further in the July, 1961, issue of the *General Radio Experimenter*. That same year, General Radio brought out its Type 1232-A Tuned Amplifier and Null Detector; this instrument used Hall's tunable notch filter as a feedback element around an amplifier to produce a continuously tunable, narrow-band audio amplifier. General Radio still markets this instrument for about $700.

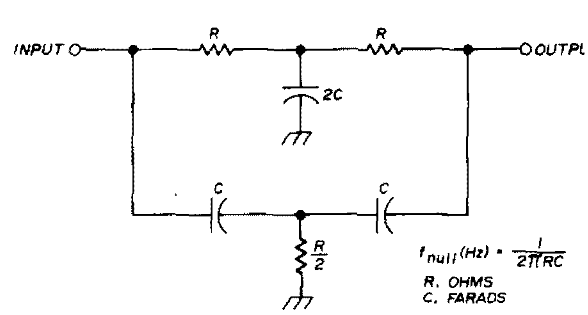


fig. 1. The twin-T notch filter is difficult to tune because three components must be varied simultaneously.

In the October, 1969, issue of *EEE Magazine*, Ralph Glasgal, in his article, "Tunable RC Null Networks," referred to the circuit as a "bridged differentiator." These articles are the only ones on this circuit of which I am aware. The following sections describe my own investigations.

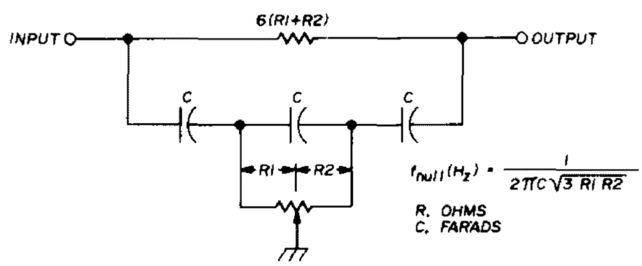


fig. 2. Basic circuit for a tunable RC notch filter which requires only one variable resistor.

the circuit

Fig. 2 shows the basic configuration of Hall's tunable notch filter. The tuning pot is composed of R1 and R2, R1 being the resistance between one end of the pot and the wiper, and R2 the resistance between the other end of the pot and the wiper. The other resistor in the circuit must have a value equal to six times that of the pot. All three capacitors have the same value, which simplifies things if you want to match them for best performance.

The null frequency equation in **fig. 2** may be solved for different values of R1 and R2, and the results plotted to show how the notch frequency changes as the wiper is moved from one end of the pot to the other. **Fig. 3** shows the result if the pot is assumed to be linear. When the wiper of the pot is at its mid-position (50 per cent of rotation) the notch frequency has its lowest value, and very little change in null frequency occurs in the middle section of the pot. As the wiper approaches either end of the pot, however, the null frequency begins to increase quite rapidly and would theoretically become infinite at the ends of the pot.

From the curve of **fig. 3** it can be seen that, in a practical tuned notch filter, the tuning pot need only cover a relatively small percentage of the range shown. In other words, if the actual pot were used with fixed series resistors so that the pot covered only the range from 1 to 10 per cent, the relative frequency range would be from 10 to 3.3, or a 3 to 1 frequency range. **Fig. 4** shows a schematic of how this would be implemented. Once the value of the pot has been selected, the values of the

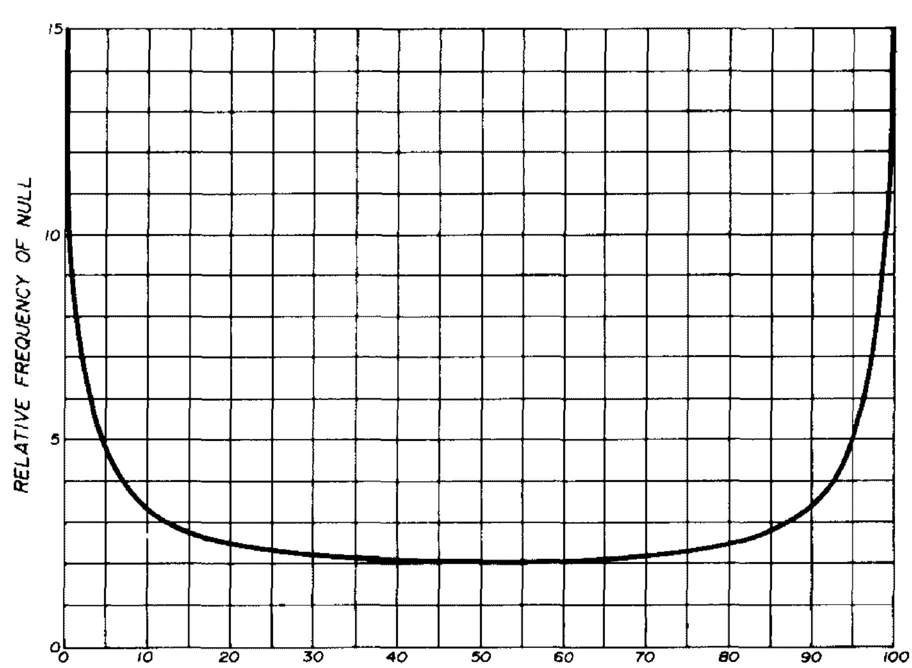


fig. 3. Relative notch frequency of the circuit shown in fig. 2 vs the rotational position of the potentiometer wiper.

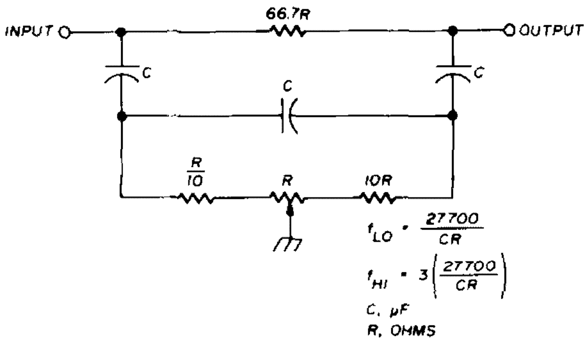


fig. 4. Relationship of resistance values for a 3:1 notch-filter tuning range.

three resistors may be calculated from the relationships shown.

It can be shown that the lowest notch frequency is given by

$$f_n = \frac{27700}{CR} \qquad (1)$$

where f_n is in Hertz, C is in microfarads, and R is the resistance of the pot in ohms. The highest frequency to which the notch may be set is equal to three times the value given by **eq. 1**.

practical examples

As a starting point, I had on hand a 20k pot and some 5 per cent, 2000 pF dipped-mica capacitors. These components would give a lowest notch frequency of

$$f_n = \frac{27700}{(0.002)(20 \times 10^3)} \approx 693 \text{ Hz}$$

The highest notch frequency should be three times this value or about 2079 Hz.

Using the resistor relationships of **fig. 4**, the circuit design was completed as

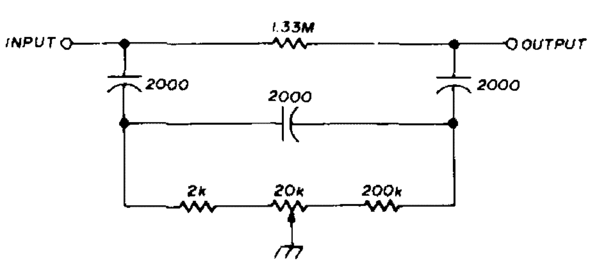


fig. 5. Notch filter which tunes from 693 to 2079 Hz. Performance of this circuit is plotted in fig. 6.

shown in **fig. 5**. It was necessary to connect resistors in series to arrive at some of the values shown, and the pot had a tolerance of at least ±10 per cent. **Fig. 6** shows the response of this filter with the tuning pot set at each end of its rotational range. Depth of the notch varies, but is greater than 50 dB in all cases; at 1160 Hz, the notch depth is 54 dB. Actual notch frequencies at the ends of the pot match the calculated values reasonably well, and the tuning range is slightly over 3 to 1.

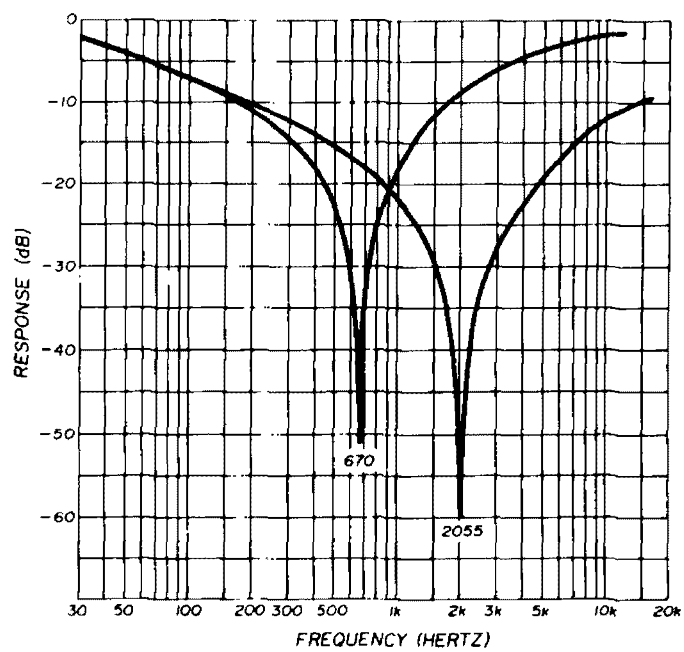


fig. 6. Frequency response of the notch filter of fig. 5 with the potentiometer set at each end.

The response curves of **fig. 6** point out the value of tuning capability in notch filters. Because the notch is so sharp and narrow at the point of maximum attenuation, it is difficult to get the peak right on a particular frequency with a filter built with fixed-value components. With a tunable notch, however, it's a simple matter to set the notch right where you want it.

An example of a tunable notch filter for 60 Hz is shown in **fig. 7**. Such a filter can often be useful in high gain audio and sub-audio circuits to mini-

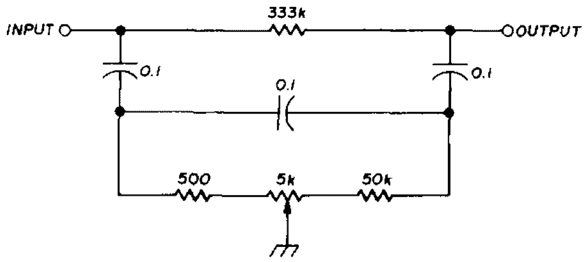


fig. 7. Tunable notch filter for 60 Hz can be used to minimize hum pickup from the ac line (circuit tunes from 40 to 120 Hz).

mize the effects of stray pickup from the ac power line. This circuit tunes from about 40 to 120 Hz.

In the examples above, the tuning ranges can be easily changed by multiplying or dividing the capacitor values by a factor. For example, doubling the capacitor values of fig. 5 would result in a tuning range of about 335 Hz to 1028 Hz, while cutting the capacitor values in half would give a tuning range of about 1340 to 4110 Hz.

optimizing notch depth

Greatest notch depth will result when the ratios of component values approach the exact theoretical design values. In this regard, there are two requirements: the capacitors should all be exactly equal in value, and the large-value resistor connected from input to output should be exactly six times the resistance in the variable-resistance branch of the network.

Capacitors may be matched by measuring them on a capacitance bridge or a direct-reading capacitance meter and selecting those most nearly equal in value. An easy way to optimize the resistance ratio is to replace the resistor connected between the input and output with a trimpot and fixed resistor, or simply a trimpot, as shown in fig. 8; the trimpot can then be set for the deepest notch. Since changing this resistance also affects the notch frequency, it will be necessary to repeatedly adjust first the trimpot and then the tuning pot until the notch can no longer be improved.

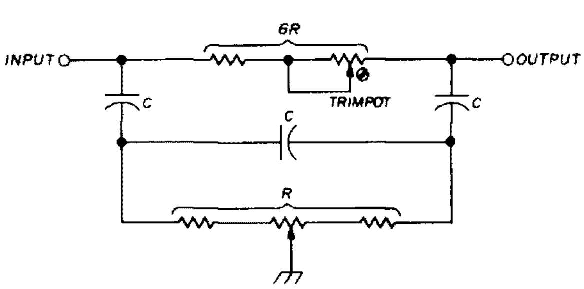


fig. 8. Notch depth may be optimized by using a trimpot to set the series resistance to exactly 6R. Notch depth of the 60 Hz filter in fig. 7 was increased from 44.5 to 57 dB using this simple technique.

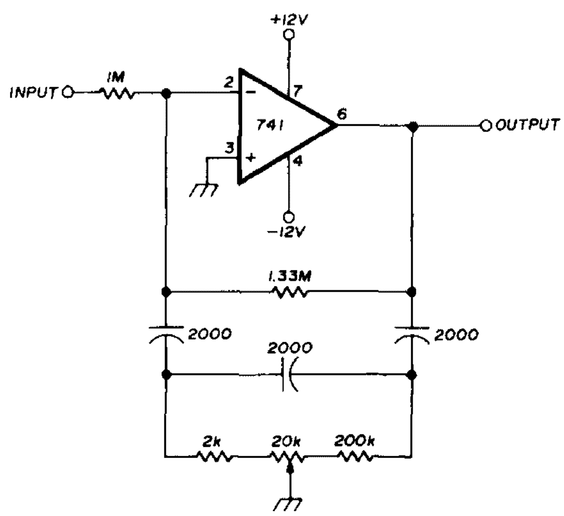


fig. 9. Tunable bandpass audio amplifier uses RC notch circuit as the feedback element. Tuning range of this circuit is 700 to 2000 Hz. Frequency response at 1000 Hz is plotted in fig. 10.

Once the trimpot has been set, however, it needs no further adjustment when the tuning pot is set to another frequency.

As an example of what this optimization can mean, the 60 Hz notch filter of fig. 7 was first built as shown with unmeasured ceramic disc capacitors and 5 per cent resistors; at 60 Hz it had a notch depth of 44.5 dB. By selecting capacitors with equal values and replacing the 333k resistor with a 500k trimpot, I was able, by careful adjustment, to increase the notch depth to 57 dB.

tuned amplifier

The tuned notch filter can be used as

the feedback element with an op amp to produce a continuously tunable narrowband audio amplifier. I used the circuit of fig. 5 in conjunction with a 741 op amp to build a tunable bandpass audio amplifier which will tune from about 700 to 2000 Hz. This circuit is shown in **fig. 9**. With the tuning pot set for a center frequency of 1000 Hz, the 3 dB bandwidth is 23 Hz, the 6 dB bandwidth is 39 Hz, and the 10 dB bandwidth is 68 Hz. At 1000 Hz, the voltage gain of the circuit measured 36 dB.

Fig. 10 shows the frequency response of this circuit when tuned to 1000 Hz. High-frequency rolloff is quite good, being about 43 dB down at 2000 Hz, so this circuit can convert a 1000 Hz square wave into a very nice sine wave. Low-frequency response flattens out, however, to a value determined by the ratio of the 1.33 megohm resistor to the 1 megohm input resistor. Some highpass filtering would improve the low-frequency skirt considerably, however. No attempt was made to optimize this notch circuit as described previously because the bandwidth is quite narrow as it stands.

Although I used plus and minus 12-volt power supplies for the 741 op amp, two 9-volt batteries should work okay; battery drain should be about 1 mA or less with a 2000 ohm load.

I have found that trying to increase the tuning range of this amplifier by increasing the value of the tuning pot and decreasing the value of the 200k resistor results in self-oscillation when the pot is set near its low frequency extremity.

This article contains everything I know at the present time about this tunable notch filter circuit and its applications. I hope I have presented enough information for those interested to make good use of this valuable but little known circuit. All data was taken using a 600-ohm audio generator and an ac vtvm with an input resistance of 10 megohms. If readers have questions or can further enlighten me on these circuits, I would be pleased to hear from them.

ham radio

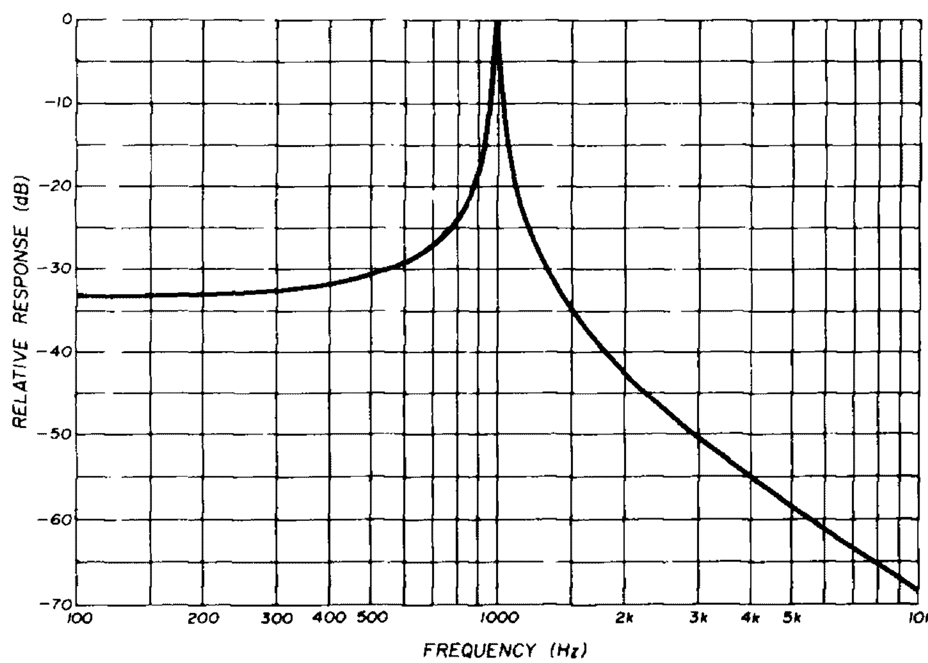


fig. 10. Frequency response of the tunable bandpass audio amplifier when tuned to 1000 Hz. Bandwidth at −3 dB points is 23 Hz; at −30 dB the bandwidth is approximately 773 Hz.

optimization of the phase-locked RTTY terminal unit

P. Edward Webb, W4FOM, Digicom, 22 Hartshorn Avenue, Worcester, Massachusetts 01602

The original design concept of the phase-locked loop RTTY terminal unit included totally automatic and universal operation requiring a minimum of adjustment. The initial design would copy any afsk shift between 150 and 100 Hz, operate at data rates up to 110 baud, and automatically track a drifting signal.[1] Another publication described a modified autostart system that incorporated an anti-CW feature.[2]

Since the original article was published there have been two distinct developments in amateur RTTY operation:

1. Wide shift (850 Hz) has almost completely disappeared from the high-frequency amateur bands with the exception of some operation on 80 meters (and, of course, wide shift is still the rule on two-meter fm).

2. The continued use of 60 wpm for 95 per cent of amateur RTTY operation, even with the availability of 75- and 100-wpm speeds.

This latter development is undoubtedly because of the vast amount of 60-wpm teleprinter machines available and the fact that most RTTY operators can't type more than 60 wpm.

With these two developments in mind, and in response to requests from readers, I've attempted to optimize the detector circuit to decrease the required minimum input-signal-to-noise ratio of 6 dB (for 99 per cent correct copy). In consideration of users having commercially made printed-circuit boards,[3] one constraint was placed on the modifications: no printed-circuit changes would be necessary — only parts substitutions, deletions, and parts value modifications would be used.

modifications

For the phase-locked loop detector (**fig. 1**) changes are shown circled. In the original loop, the lowpass filter capacitor, C7, was selected to give a loop bandwidth of approximately 400 Hz. This value was selected in compliance with Shannon's rule for operation at the highest data rate of 110 baud: 110 baud ÷ 2 = 55 Hz signaling rate; 55 Hx × 7 = 385 Hz, or the required minimum loop bandwidth. But for 45-baud (60 wpm, 5 level), a 22.5-Hz signaling rate results, thus requiring a loop bandwidth of only 157.5 Hz. A 180-Hz loop bandwidth was used because 0.22 μF was the nearest standard value capacitor to the design value for C7. To complement the preceding change, the phase-locked loop output RC ladder filter cutoff frequency was also lowered by increasing the value of C8, C9, C10 and C11 from 0.01 to 0.022 μF.

The ssb position of the mode switch, S1, is now used for 170-Hz shift, and the vco frequency-set pot, R5, is adjusted for a natural vco frequency of 2210 Hz at TP-2 with no audio input applied to the TU. This new frequency setting allows the vco to swing symme-

trically to either side of its natural frequency, thus reducing the probability of the loop unlocking on noise.

The *normal* position of the mode switch, S1, is used for 850-Hz shift and its pot, R6, is adjusted for a natural vco frequency of 2550 Hz at TP-2, as in the original article. Other shifts, such as 425 or 85 Hz, can be copied simply by adjusting the receiver frequency so that the shift is centered on the appropriate vco natural frequency.

In the tracking comparator, an RC network consisting of R11 and C12 is

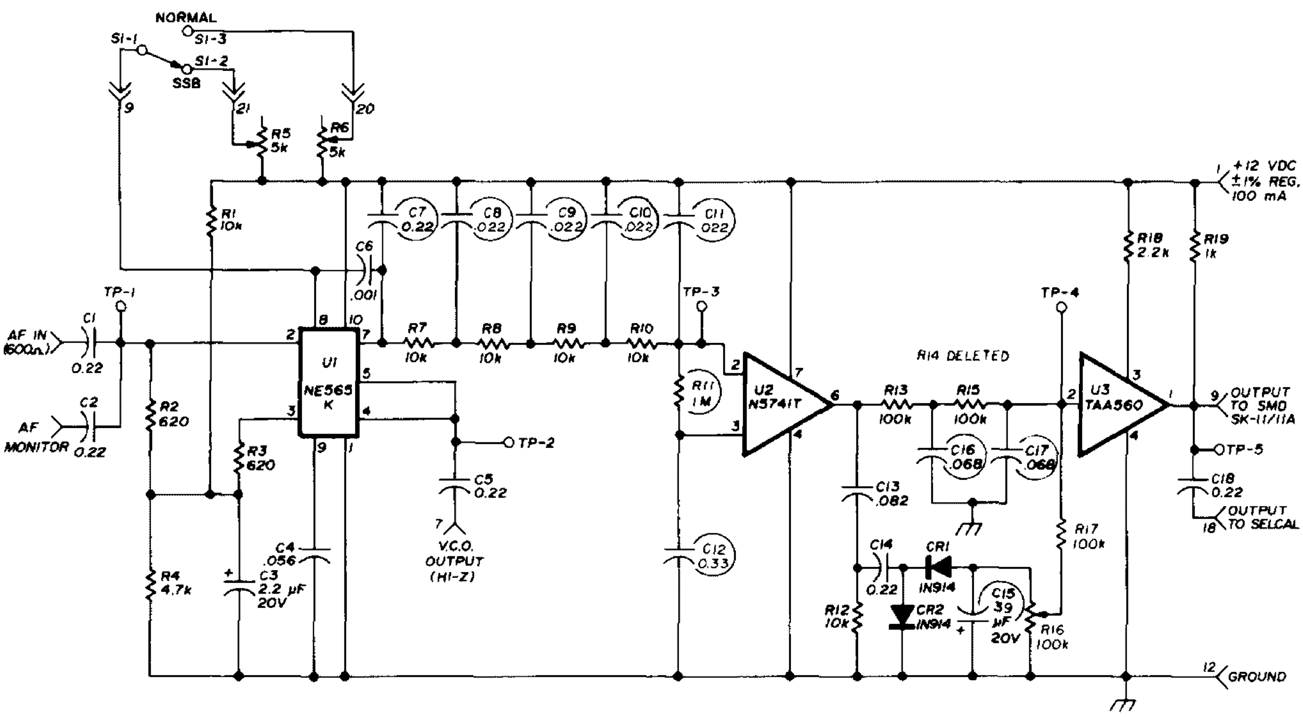


fig. 1. Circuit modifications to the phase-locked loop RTTY terminal unit for improved performance. New circuit values are circled. All resistors are 5%, ½ watt unless otherwise specified.

used to produce a signal-derived reference voltage for the noninverting input of U2. The RC time constant was selected to be 2.5 times that of the longest *space* condition at 60 wpm, which is the *blank* key having a *space* duration of 132 milliseconds. Accordingly R11 and C12 were selected to give a time constant of 330 milliseconds. However, the leakage path through electrolytic capacitor C12 limited the minimum detectable shift to approximately 150 Hz. Thus C12 has been changed to a 0.33 μF mylar, and R11 has been changed to 1 megohm to maintain the same time constant. To take advantage of the higher impedance presented by this RC network, U2 has been changed from a N5741T to a NE536T, an fet input op amp. The minimum detectable shift was measured at 40 Hz with an input-signal-to-noise ratio of 10 dB and a noise bandwidth of 3 kHz.

The tracking comparator output lowpass filter cutoff frequency has been decreased to 30 Hz to optimize the section for the 22.5-Hz signaling rate used with 60 wpm, 5-level RTTY. To accomplish this frequency change, capacitors C16 and C17 were increased from 0.033 to 0.068 μF. Also affected was resistor R14, which was deleted to produce a higher voltage to the input to the Schmitt trigger, U3, so that its operation on marginal signals would be improved. The noise rectifier in the noise squelch circuit was changed. The output filter capacitor, C15, was in-

creased from 10 to 39 μF to provide a longer time constant, which reduced the effects of static crashes.

experimental procedure

The phase-locked loop[2] TU was connected to a test setup (**fig. 2**). A Wavetek model 134 function generator was used to generate a 22.5-Hz square wave to simulate a 60-wpm, 5-level RTTY signal. This signal was fed through an attenuator to the voltage-controlled generator input of a Wavetek model 132 function and noise generator. This latter instrument could simultaneously generate both a signal as well as random noise in a front-panel controlled, calibrated ratio. In these tests the signal used was an afsk sine wave shifted 170 Hz between 2125 and 2295 Hz, and a random noise sequence of $2^{20} - 1$ was used at a maximum noise clock frequency of 48 kHz, thus ensuring that a random noise sequence could not repeat in a time period of less than 21.85 seconds.

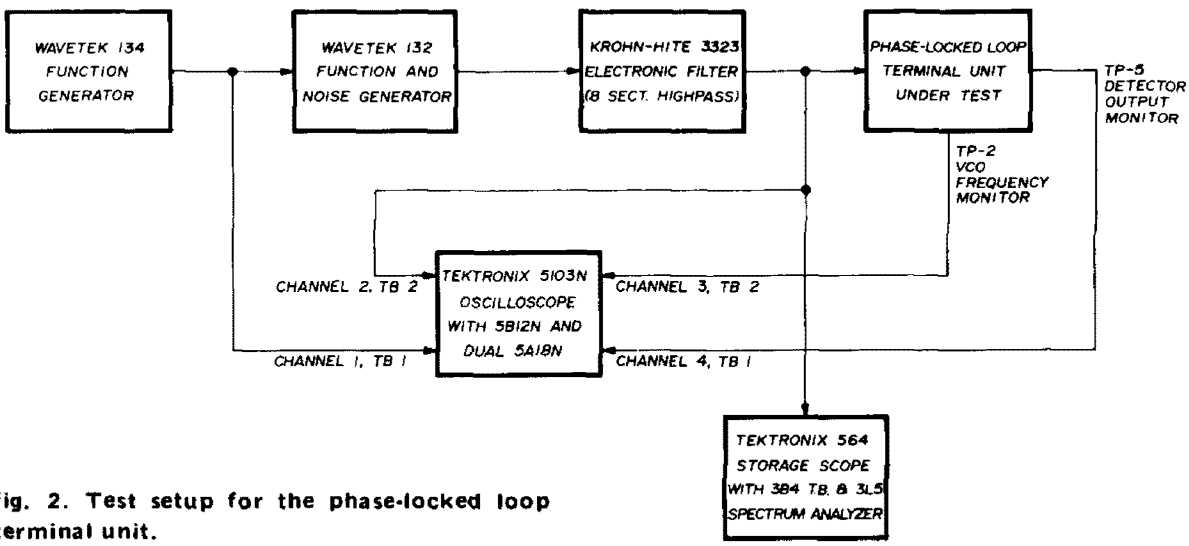


fig. 2. Test setup for the phase-locked loop terminal unit.

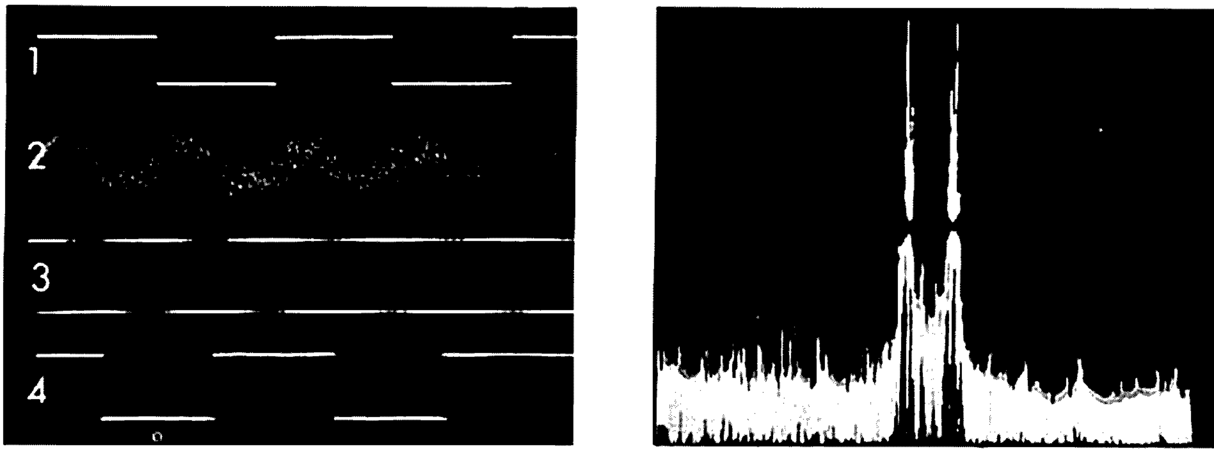
Noise clock frequencies of 48, 45 and 39.3 kHz were used to produce upper noise bandwidths of 3, 2.81, and 2.46 kHz so that the signal-plus-noise bandwidth was always centered on 2210 Hz, which is the shift midpoint and the natural frequency of the phase-locked loop[2] vco. A lower bandwidth limit of 10, 1610, and 1960 Hz was imposed by the use of a Krohn-Hite dual electronic variable filter with both sections connected in series for highpass operation. These instrumentation methods gave simulated receiver bandpasses of 3000, 1200 and 500 Hz respectively.

All tests were started at a 10 dB s/n ratio, which was then reduced until the output of the Schmitt trigger (**fig. 3**, Channel 4) exhibited excessive jitter or failed to follow the afsk keying signal (**fig. 3**, Channel 1). This s/n ratio was considered to be the minimum usable input signal for a given simulated receiver bandpass. The test results were: 0 dB minimum s/n for a simulated receiver bandpass of 3 and 1.2 kHz. For 500 Hz, it was -3 dB.

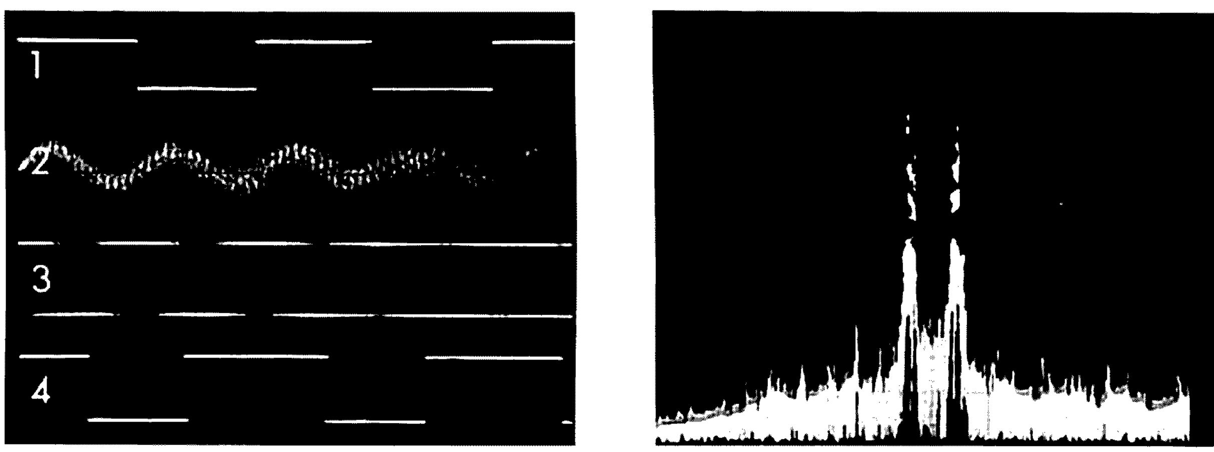
conclusion

Figs. 3A, 3B and **3C** illustrate the input and output waveforms for the test setup and TU with input s/n ratios of zero and -3 dB and simulated receiver bandpasses of 3, 1.2, and 0.5 kHz. These figures represent a 6-dB improvement for the worst case and a 9-dB improvement for the best case over the original design.

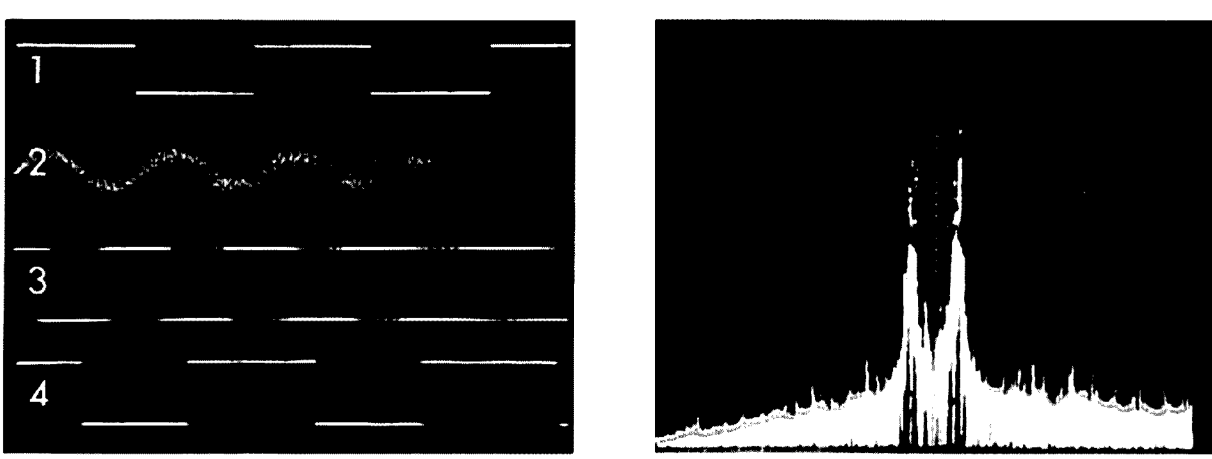
Comments, contribution of ideas, and criticism are welcome from users of phase-locked loop terminal units.



A. 3-kHz noise bandwidth; signal-to-noise ratio, 0 dB.



B. 1.2-kHz noise bandwidth; signal-to-noise ratio, 0 dB.



C. 500-Hz noise bandwidth; signal-to-noise ratio, −3 dB.

Center frequency, 2200 Hz; dispersion, 200 Hz/cm; sweep speed, 5 sec/cm; vertical sensitivity, 200 mV/cm.

fig. 3. Input and output waveforms from test setup show improvements of 6 and 9 dB over original design for worst- and best-case conditions.

Channel 1: afsk keying voltage; horizontal, 10 ms/cm; vertical, 50 mV/cm.
Channel 2: signal + noise at 3 kHz bandwidth; horizontal, 0.2 ms/cm; vertical, 100 mV/cm.
Channel 3: PLL vco at TP-2: horizontal, 0.2 ms/cm; vertical, 10 V/cm.
Channel 4: Schmitt trigger output: horizontal, 10 ms/cm; vertical, 10 V/cm.

acknowledgement

I'd like to express my gratitude to the Wavetek Corporation for making a Model 132 VCG/noise generator available for this and other research work in the biomedical field. The work reported here was greatly simplified by the use of this instrument.

references

1. P. Edward Webb, W4FQM, "A Phase-Locked Loop RTTY Terminal Unit," *ham radio*, January, 1972, page 8.
2. P. Edward Webb, W4FQM, "An Anti-CW Autostart System For the PLL TU," *73*, May, 1972, page 39.
3. Digital Communications Corp., 185 Devonshire, Suite 900, Boston, Massachusetts 02110.

ham radio

toroidal coil inductance

Toroidal inductors are being used more and more in amateur communications equipment because of their small size, high Q and self shielding properties. Michael Gordon's article* makes it relatively easy to determine the number of turns to wind a coil on a particular core. He includes numerous core constants and a helpful equation necessary to achieve a specific inductance. **Fig. 1** contains much of that information in graphical form and was developed for builders who tend to shy away from using mathematical formulas. To determine the required number of turns on a particular toroid core, simply locate the desired inductance, L, on the vertical axis, draw a horizontal line through L to where it intersects the appropriate core line, and read off the number of turns, N, on the horizontal axis, beneath the intersection point.

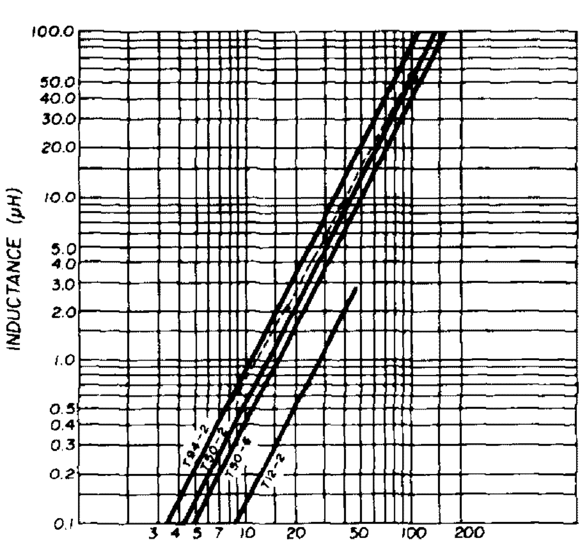


fig. 1. Inductance vs number of turns for four popular Amidon toroidal cores. Dotted line shows actual measured inductance values for a T50-2 core; discrepancy from calculated values is small.

The graph of **fig. 1** can be expanded to include more core types and a greater number of turns if desired. Fortunately, Gordon's expression $N = K\sqrt{L}$ plots as a straight line on log-log graph paper; therefore, by using paper with more cycles, a more complete nomograph can be obtained. I constructed the original graph by rearranging the above formula into the form $L = N^2/K^2$, where K is one of the core constants, and then solved the equation for L when N is 20 and 100. This gives two points which can be connected by a straight line.

The dashed line shows actual measured values for an Amidon T50-2 core. Four test coils were wound using number-26 (0.4mm) enamel covered wire. A commercial Q meter was used to measure Q, and the inductance was calculated from the resonant frequency formula. The small discrepancy noted between the measured and calculated inductance values could have been due to connecting lead length and Q-meter accuracy.

Charles Miller, W3WLX

*Michael Gordon, WB9FHC, "Calculating the Inductance of Toroids," *ham radio*, February, 1972, page 50.

single sideband speech splatter

An informative discussion of nonlinear ssb operation and what to do about it

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If you've listened on the low-frequency end of the 20-meter phone band when two or three rare DX stations are coming through, you've probably noticed a steady, noise-like signal between the DX stations. Under good conditions the noise may be S2 or S3; more likely it's S9 or more. This steady noise seems to come from no particular station; in fact it seems to emanate from all stations of the band. The noise is caused by splatter.

When only one station is causing splatter, it's a good bet the operator will be told his transmitter is at fault. Sometimes this brings an apology and an immediate resolution; often it brings a protestation of innocence based on the equipment in use: "My so-and-so off-the-shelf rig has super automatic level control; it can't be me." A cop-out response? Perhaps not. The fellow in question may not understand the causes and cure of splatter.

This article presents some hints on the causes, cure, and source of splatter interference and what can be done about it to help alleviate the noise problem in the congested amateur phone bands.

source of splatter

Splatter comes from a single source; in fact a rather simple one. There is a relationship between the way a signal changes and the bandwidth it occupies. The relationship is easily described mathematically, but this does not lead to a good understanding of the process. For this discussion let's look at some

special types of signals and use these as analogies.

A simple example of a signal with a high rate of change is a pulse of width Δt, corresponding to a dot of a CW signal or to an ideal radar pulse. With respect to time, the signal is off for a long period, then on at constant strength for a period Δt, then off again for a long period. With respect to frequency, the relative strength of this signal varies with frequency; its spectrum is shown in **fig. 1**. Most of the signal energy is contained in the bandwidth defined by $f = 1/\Delta t$.

For example, suppose the pulse represents a dot with a width of 50 milliseconds and a dot speed of 24 words per minute. Most of the energy will be contained within a band 20-Hz wide. If the pulse is from a radar, having a width of 0.2 microsecond, most of the signal energy will be contained within a band 5-MHz wide. On a relative basis, the shape of these two signals is the same.

Both of these signals have additional peaks extending outward from this central band, as shown in fig. 1. These additional components are the source of key clicks on CW signals and the typical buzz of a radar signal when a receiver is tuned well away from center frequency.

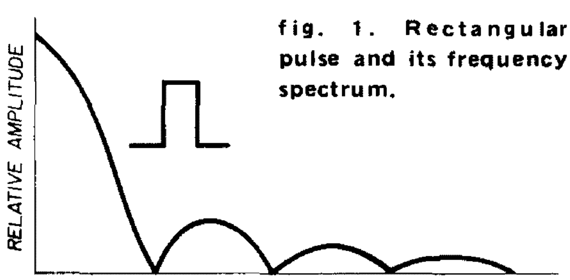


fig. 1. Rectangular pulse and its frequency spectrum.

The components extend far from the center frequency.

For a perfectly rectangular pulse, the magnitude of the components is defined by the relationship (sin X)/X. This relationship is called the aperture function and is the envelope of the spectrum amplitude for a single square pulse. The X, in this case, is $\pi T_o f$, where T_o is the pulse duration (seconds) and f is the frequency (Hz) at which it is desired to know how much signal amplitude exists. If the pulse duration is 0.1 second and you want to know how much splatter

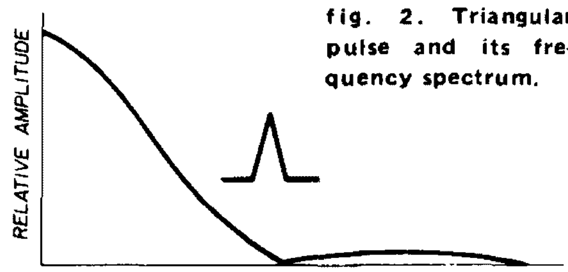


fig. 2. Triangular pulse and its frequency spectrum.

exists at, say, 100 Hz from center frequency, then

$$\pi(0.1)(100) = 10\pi$$

and

$$\frac{\sin 10\pi}{10\pi} = 0$$

There is no splatter at all at exactly 100 Hz from center frequency. However, if you ask about splatter 5 Hz from center frequency,

$$\pi(0.1)(5) = 0.5\,\pi$$

and $(\sin \pi/2)/\pi 2 = 1/\pi 2 = 2/\pi = 0.636$, or about two-thirds the signal you'd hear exactly on frequency.

The influence of this idealized signal extends over the entire band, from dc to infinitely high frequencies. The change in signal strength with frequency is slow. For example, if the maxima have a strength of 100 mV at 10 kHz from the center of the CW signal, they will have a strength of 50 mV at 20 kHz from the signal, 25 mV at 40 kHz, etc.

For a given pulse width and amplitude, the only way we can change the signal spectrum is to change the rate of change of the signal or rise time. This affects the strength of components far

from the center frequency of the signal without changing the main part of the signal very much. For example, suppose the pulse were changed from a perfect rectangle to the triangle of **fig. 2**. The energy distribution is now described by the relationship $[(\sin X)/X]^2$, but the

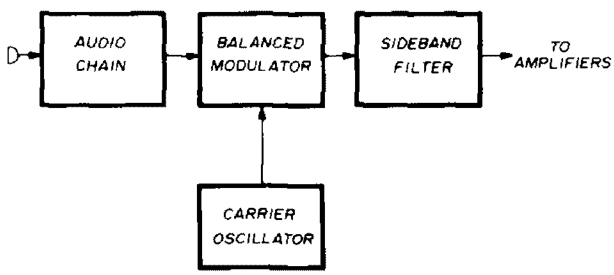


fig. 3. Block diagram of the basic filter-type ssb transmitter.

peak amplitudes far from center frequency decrease twice as fast as before. If the maxima are 100 mV at 10 Hz from center frequency, they will be 25 mV at 20 kHz; 6.25 mV at 40 kHz, etc. The energy very close to the center frequency is identical for the triangle and the rectangle. Other wave shapes can show greater reduction of the "far-out" components.[1]

This modification of wave shape and its effect on far-from-center frequency components is the principle of the key click filters used in CW. Note that the shaping *can't* stop key clicks – *all* it can do is reduce their strength and, in particular, reduce the strength of the components relatively far from the signal. For the 20-wpm signal, clicks beyond a few hundred Hz can be brought to the noise level or below.

The important point of this discussion is that any signal that goes on and off will spread across a band of frequencies. How far it spreads and how much energy occurs at each frequency is determined by the interval between on and off, the rate of change of the signal during the on period, and the maximum amplitude the signal reaches.

ssb signal generation

Now let's apply these principles to ssb phone signals. We'll ignore some "fine structure," such as the effect of turning the transmitter on at the beginning and off at the end of a transmission. What we want to look at is how a signal supposedly contained within the speech bandwidth can appear many kHz from this nominal band: in other words, we want to know where splatter comes from. To do this let's look at how the signal is formed and processed.

Fig. 3 shows, in block diagram form, an elementary single-sideband transmitter. *Audio and carrier are* processed by the modulator to produce a double sideband suppressed-carrier signal. The modulator also produces distortion products at twice the frequency of the carrier, three times its frequency, etc. However, all of the signal outside the passband of the filter is rejected; therefore, only speech products lying between about 300 and 2400 Hz, plus the small amount of distortion that lies within this range, are passed on to the following amplifiers. At the output of

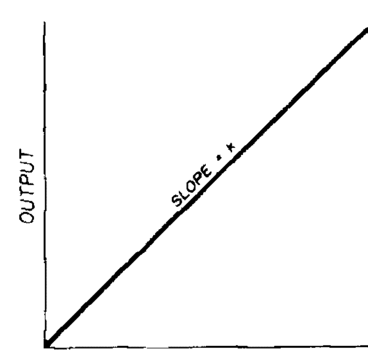


fig. 4. Ideal amplifier has linear relationship between input and output.

the filter the signal is free of widespread spurious components: it is splatter-free.

This nearly ideal signal is then amplified and finally transmitted. If the amplifiers are linear, no energy will exist outside the passband. That is, the amplifier output signals must bear a direct relation-

ship to the input signals. Each amplifier must follow the law

$$e_{out} = k(e_{in})$$

The transfer function of the amplifier must look like fig. **4**. Unfortunately, most amplifiers don't attain this ideal relationship and this is where the trouble comes from. In looking at this, let's again remember we're not interested in "fine structure;" that is, distortion close to the transmitted signal or regeneration of the rejected sideband, or things of that kind.[2] We want to know the *source* of splatter.

splatter due to flat-topping

One characteristic of practical amplifiers is that they saturate. The output reaches some maximum level then doesn't increase further. Instead of being linear they have the relationship shown in fig. **5**. The effect on a typical signal is shown in **fig. 6**; it becomes amplitude-limited and is usually called flat-topped. The effect of flat-topping is the same as if the transmitter were putting out an ideal signal plus a

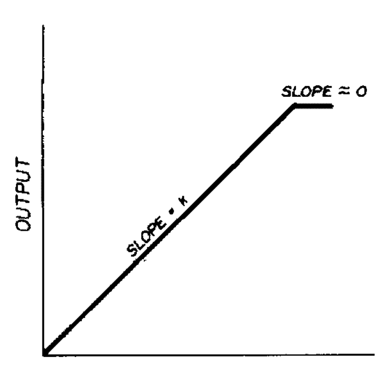


fig. 5. Input-output relationship of an amplifier which saturates.

negative-going pulse that cancels the top of the signal (fig. 7). The transmitter is now transmitting its intended signal and is also transmitting a radar-like pulse. Flat-topping generates splatter. Since speech waveforms last for relatively short periods, and since only the tip of the wave is cut off, the radar-like pulse formed by peak clipping has a very short duration. Therefore, it spreads a long way from the desired signal and repeats at each voice peak, generating the sound amateurs call "buckshot".

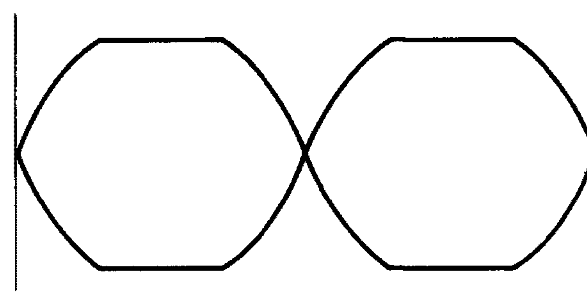


fig. 6. Ssb two-tone test with saturation.

Note that the signal close to the center frequency is perfectly good. Don't ask the station you are working if you are splattering; he can't tell since he hears a perfectly good signal. It's the station well removed from the center frequency who hears the splatter.

splatter due to bias-off

Another characteristic of practical amplifiers is the tendency to ignore small signals. For a small input they show no output for a time. The transfer function tends to look as in **fig. 8**, which causes the bottom end of the signal to be clipped. Again, this is equal to a normal signal plus a distortion term. In audio amplifiers, where push-pull operation is common, this clipping is called crossover distortion. It also exists in single-ended amplifiers and is the result of improper bias adjustment – the bias is set too high – often in an attempt to reduce the tube idling current.

This form of distortion is less likely to cause trouble than that from positive peaks, since only a small part of the signal is affected. However, evidence indicates that the importance of this form of distortion has recently in-

creased as power companies lower line voltage or allow poorer line voltage regulation. These acts reduce the plate voltage (unregulated) but usually don't affect the bias voltage, since it is regulated. Effectively, the reduced line voltage increases the bias on the tube and therefore increases this type of distortion.

splatter due to parasitics

Another characteristic of practical amplifiers is a tendency to oscillate, often at a frequency far from that at which they are designed to work. Sometimes this parasitic oscillation starts at a specific plate voltage and dies out when the voltage increases beyond this level or when it decreases below this level again. The effect on the input-output relationship can be as in **fig. 9**. Either of these oscillations will create splatter as the parasitic turns itself on and off. Very similar effects can appear if the amplifier is merely unstable instead of oscillatory. This action causes the amplifier output to increase beyond its normal value, then level off, or perhaps even drop back to normal.

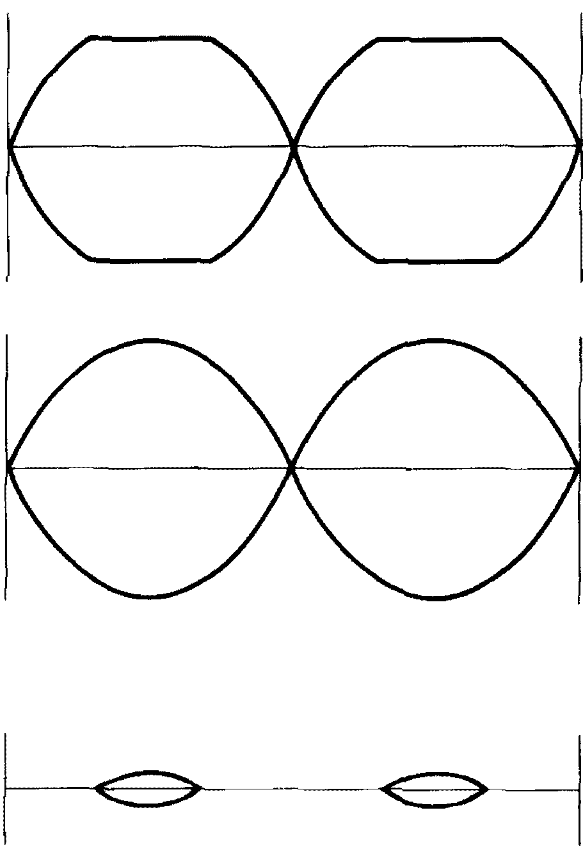


fig. 7. Flat-topping as the sum of an ideal signal plus pulses.

Problems of these types are very common when amplifiers are first tuned

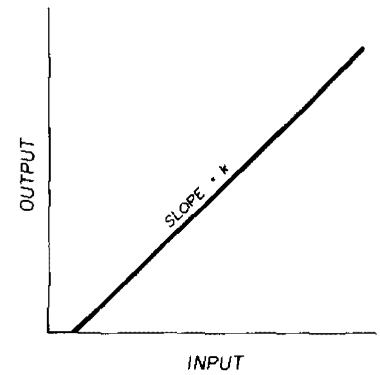


fig. 8. Input-output relationship of an amplifier with crossover distortion.

up. They may arise again because of components that become defective or change values: more commonly, they arise when an amplifier isn't adjusted properly or if it's operated with voltages appreciably different than those for which it was adjusted. It's likely that the really bad cases of splatter are due to such parasitic oscillations or instability.

splatter due to level control

Let's go back to the flat-topping situation, which is the most common problem. A number of automatic means of preventing flat-topping have been developed. Most work by detecting a particular level of signal at the final amplifier and feeding the signal back to an earlier amplifier to reduce the gain. This scheme is called automatic level control, ALC, or overdrive detection. The intent of these systems is to reduce the value of k in the linear input-output relationship, $e_{out} = k\,(e_{in})$, to the point where linear operation is maintained — by keeping all peaks below the overload point.

Unfortunately there can be some problems with these automatic circuits. It takes a finite time for the overload signal to be detected, fed back, and for it to reduce the gain of the earlier stage. This time delay can also introduce a problem, as shown in **fig. 10**. The relatively sudden decrease from the initial value of k to a second value is just as effective in producing splatter as any other change of equal rate and magnitude. An associated splatter source may be the click of the vox relay pulling in: it's gone before the ALC circuits can act, but it can still produce splatter.

Some transmitters incorporate circuits to prevent these sudden changes from occuring. There may be several gain-controlled stages, or several control sources, with different time constants. These complex circuits definitely help prevent the rapid changes shown in **fig. 10**; however, even they can't cope with the results of a wide-open gain control. The transmitter struggles zealously but just can't prevent splatter from occurring.

splatter due to the receiver

To be complete, we should note that receivers also include some linear stages, and can have almost identical problems. Problems arise if these stages are overloaded, unstable, or if the automatic gain control circuit time constants don't match the rate of change of a particular signal being received. Splatter reports can be generated in the receiver, so that a false report is perfectly possible. If you're told you're splattering, it's in order to ask about the precautions against receiver overload, assuming your transmitter is operating linearly.

detecting your own splatter

Many stations are equipped with an oscilloscope or monitor. This is by far the best way to operate, but it's necessary to be cautious. A scope won't show all splatter problems. Example: suppose two transmitters are operating. One is an ssb transmitter with two-tone modulation. The second is an a-m transmitter **100** percent modulated by the unfiltered output of a full-wave rectifier. As seen on a monitor scope, what is the difference between these two signals?

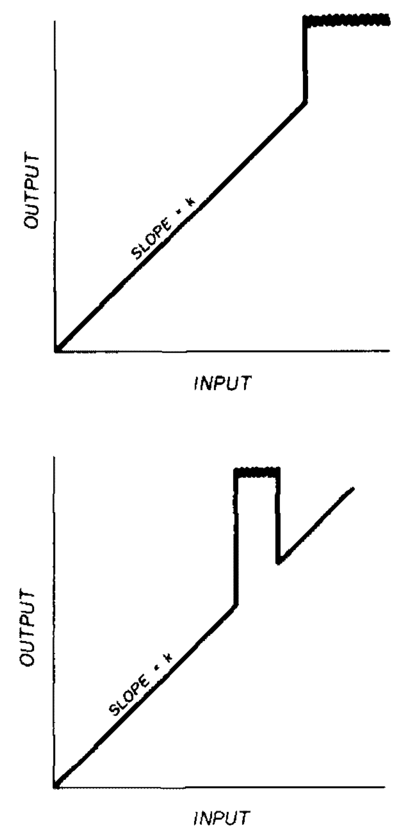


fig. 9. Input-output relationships of an amplifier with parasitic oscillations.

The answer is that, to casual inspection, there is absolutely no difference. Both look like the ideal signal of **fig. 7**. Even detailed inspection of the envelope will not show any difference. The only way a scope can detect the difference is to look at the rf waveform. When this is done, a phase reversal will be found at each zero crossing of the envelope for the two-tone test: the rf wave of the a-m signal doesn't show this reversal. This very small difference won't show up on any of the usual types of signal monitor; yet the difference as seen by outside receivers is enormous. The two-tone ssb signal is clean while the recti-

fier-output modulated a-m signal spreads to many, many times the frequency of the modulating sine wave.

Fortunately this situation is artificial but it does show that casual observation of the scope is not enough; some attention to detail is necessary. The first

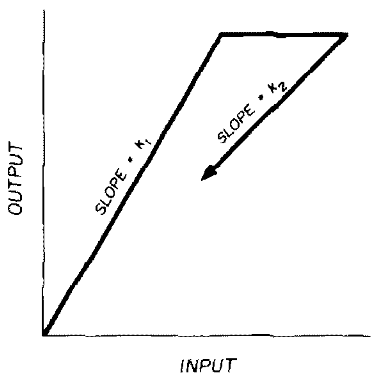


fig. 10. Input-output relationship of an amplifier with poor automatic leveling control.

point to watch for is flat-topping, usually the only item observed. Sharp peaks of the voice signal are a good indication that splatter is not occurring from flat-topping. However, splatter due to parasitics or to instability can cause a pattern that resembles, very closely, the sharp peaks of a voice pattern. It takes careful examination of the signal, and, in particular, examination of the *rate of rise* around the midpoints of the signal, to detect splatter-producing parasitics. The type of splatter that arises from sharp changes at the lower signal levels is also detectable on the scope, but also only by careful observation. The rule should be, inspect the *entire* signal, not just peaks.

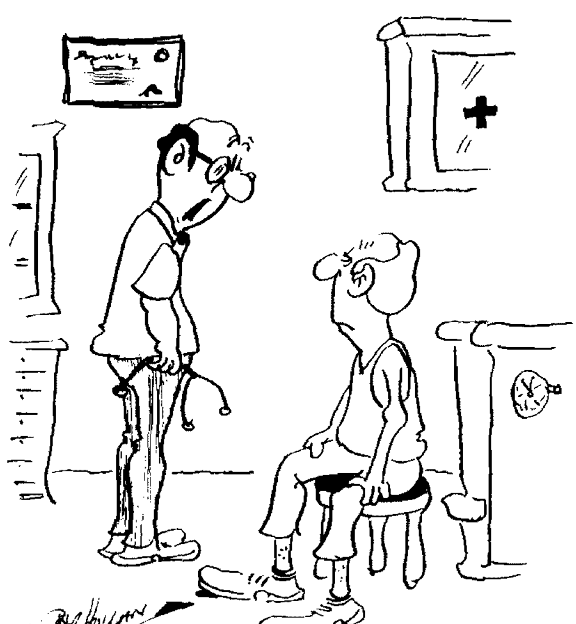
Many operators with stations using transmitters incorporating automatic level control (ALC) eliminate the scope and place their dependence on the automatic circuits. As long as the circuits are working properly and the transmitter is adjusted properly, this is satisfactory. The circuits do work if given half a chance; however, the practice of turning the gain fully open is responsible for much splatter. The gain control should be set to give ALC action as recommended by the instruction book.

Stations using transmitters without ALC and no scope have a special problem. It's almost impossible to detect splatter generation by reading the meters on the transmitter. About the only automatic way of preventing splatter is to choose an exciter that *can't* overdrive the amplifier. In most cases, this means limiting the exciter power to about one-tenth to one-twentieth the power of the final amplifier. If the final is to run 1000 watts input, the exciter should run no more than 50-100 watts input.

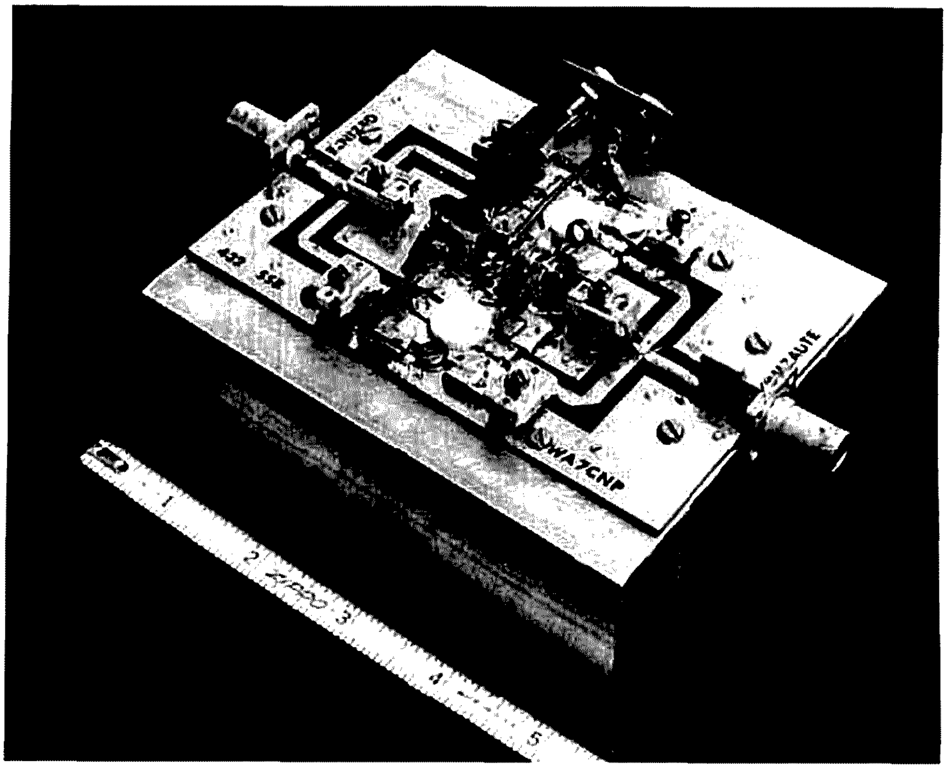
references

1. *Reference Data for Radio Engineers*, Fifth Edition, "Fourier Waveform Analysis," page 42-1.
2. Marv Gonsior, W6VFR, "Intermodulation Distortion Measurements on SSB Transmitters," *ham radio*, September, 1974, page 34.

ham radio



"Let me put it this way. If you asked me your QSA, I'd give you about a 2-3-1."



100-watt solid-state power amplifier for 432 MHz

Design and construction of a 100-watt solid-state power amplifier for linear, CW, or fm service

R. Keith Olsen, WA7CNP*

There has been a tremendous increase in ssb activity on 432 MHz since the launch of Oscar 7 in late 1974. However, many amateurs who are communicating through Oscar's 432-to-144-MHz translator are using only 5 to 15 watts ssb output on 432 MHz so they must use complex, high-gain antennas to hold a good signal through the satellite. Few 432-MHz power amplifier designs have appeared in print, and because of mechanical considerations, vacuum-tube amplifiers are difficult to build at this frequency. This article describes a two-transistor, 100-watt PEP, solid-state linear amplifier which solves many of these problems. The components used in the amplifier, for the most part, are readily available.

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The transistors used in this power amplifier are the new high reliability Motorola MRF306s. The MRF306 is a 28 volt, 60 watt, 225 to 400 MHz power transistor and is unique in that it is internally matched for broadband use. Internal matching is nothing more than extending the matching network to the actual chip inside the package. This is accomplished by incorporating the bond wires and additional MOS capacitors as tuning elements between the base lead and thr transistor die (fig. 1). The ladder network transforms the input impedance of the device in a controlled manner so it represents a low loaded Q at the frequency of use, hence the name, "Controlled Q Transistor."

Gold metalization is used inside the MRF306 to provide high reliability, ruggedness, and long life. The device can withstand a load vswr of 30:1 at *all* phase angles at 60 watts CW output. This kind of ruggedness offers burnout protection against hazards such as a forgotten antenna connector or even a mislaid screwdriver!

theory of operation

The power amplifier described here is essentially a narrowband, parallel amplifier that can be tuned from 420 to 450 MHz. For conventional ssb use, a push-pull configuration should be employed; in this case, however, both transistors are being driven at less than what they are capable, and operate in the class AB linear mode. **Fig. 2** is a spectrum display of a two-tone test I ran on the amplifier. The center frequency is 432 MHz, the two tones are spaced 500 kHz apart, and power output is 100 watts PEP. The drive level is approximately 10 watts PEP.

As you can see, the 3rd and 5th order intermodulation products are down 28 dB. IMD responses greater than −30 dB can be attained by running the amplifier at lower power levels (see **fig. 3**), but since 100 watts is intended to be the "worst-case" condition, I chose to demonstrate its response there. I might add that you should always pay very close attention to the linearity of any amplifier, especially when operating in crowded bands or when using one of the OSCAR satellites. When you operate an amplifier too close to the saturation point of the curve, you not only increase the incidence of spurious emission, your ssb signal also becomes mushy and distorted.

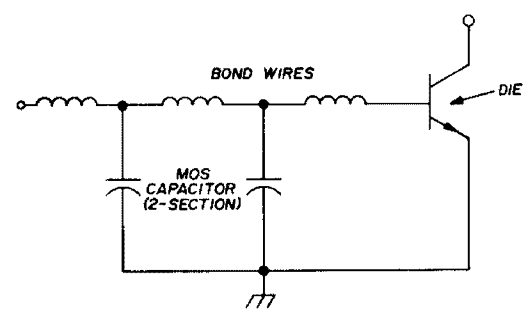


fig. 1. The MRF 306 power transistors used in this amplifier use a ladder network between the base lead and the transistor die to transform the input impedance of the device so it represents a low Q at the frequency of use.

This power amplifier can also be used on CW and fm. When I first built the amplifier I biased it into class C, and was able to drive and sustain 140 watts of output power for more than 30 minutes (I cheated a bit by using a fan because the ambient temperature of the heatsink can rise above 80°C). If you wish to use the amplifier for either CW or fm, and not on ssb or a-m, you can replace the class AB bias circuit with a ferrite bead and a four-turn inductor from the base of each device to ground as shown in **fig. 5A**. The devices can be driven further into class C by placing a 2.7 ohm, ¼ watt resistor between the inductor and ground (**fig. 5B**). This resistor is required only if you desire greater efficiency or have a greater drive level capability.

construction

The amplifier is built on 1/16 inch (1.5mm) thick double-clad G10 epoxy-

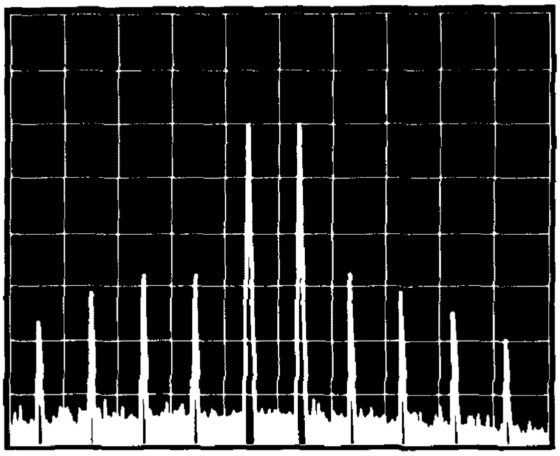


fig. 2. Spectrum display of 100-watt amplifier at 432 MHz shows 3rd and 5th intermodulation distortion products are down 28 dB referenced to 100-watt output (each vertical division = 10 dB).

glass printed-circuit board. I normally design with 1/16 inch (1.5mm) thick glass-teflon board, but at $135 a sheet it's a bit expensive for amateur radio projects. Epoxy-glass circuit board is somewhat lossy at 432 MHz, but is still quite acceptable for amateur use considering the cost differential.

However, it is imperative that you use double-sided board. It is also important that the ground on the bottom side be tied closely to the ground pads on the top side. This is accomplished by using either eyelets or plated-through holes from one side to the next. If you don't have eyelets you can do it by drilling holes through the board with a number-50 (1.8mm) drill; you then place number-18 or -20 (1.0 or 0.8mm) wire through the holes, solder both ends, and trim them off flush with the circuit board. I recommend that you make such a connection under each Unelco capacitor. As for the holes for the transistor flanges, a good drill, a file, patience, and a little elbow grease are all that is necessary.

The Unelco metal-clad mica capacitors have a dual purpose. Not only are they important to the input and output networks, they also serve as supports for the transistor leads. The Unelco capacitors are a bit expensive, but at these frequencies they cannot be replaced by anything other than porcelain chip capacitors such as those manufactured by ATC.

The transistors should *always* be screwed down to the heatsink (thermal compound underneath) *before* soldering the leads as it is possible to crack or break off the ceramic cap of the device. (The clearance holes in the flange are made for 4-40 [M3] screws.) Once this is done, the base and collector leads can be connected to the microstrip with 2 to 5 mil (0.05 to 0.10mm) thick copper or brass strap. Be sure that the width of the strap does not exceed the width of the microstrip.

The Arco variable mica trimmers work surprisingly well at uhf. Piston variables are better suited for uhf work, but here again the price tag is prohibitive. When you look at the bottom of the Arco variables you will notice that there are two small grounding tabs on each side. These tabs must be soldered to the ground plane to reduce the amount of lead inductance inherent in the capacitor. Be careful, when installing the capacitor, to make sure it doesn't short out the microstrip.

The collector and base biasing circuits are best placed between the two devices in the center of the board (see **fig. 6**). The 580 pF feedthrough capacitor, C12, is mounted on a piece of 0.7x0.4 inch (18x10mm) copper strap.

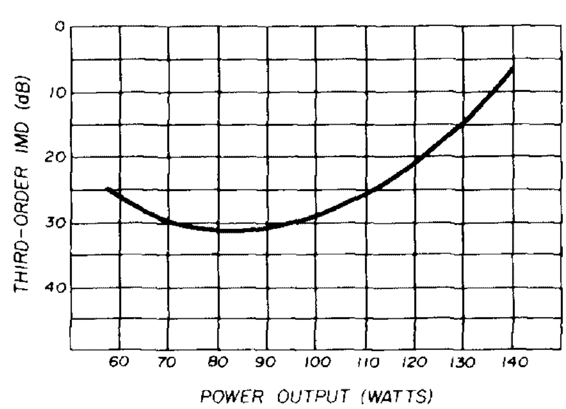
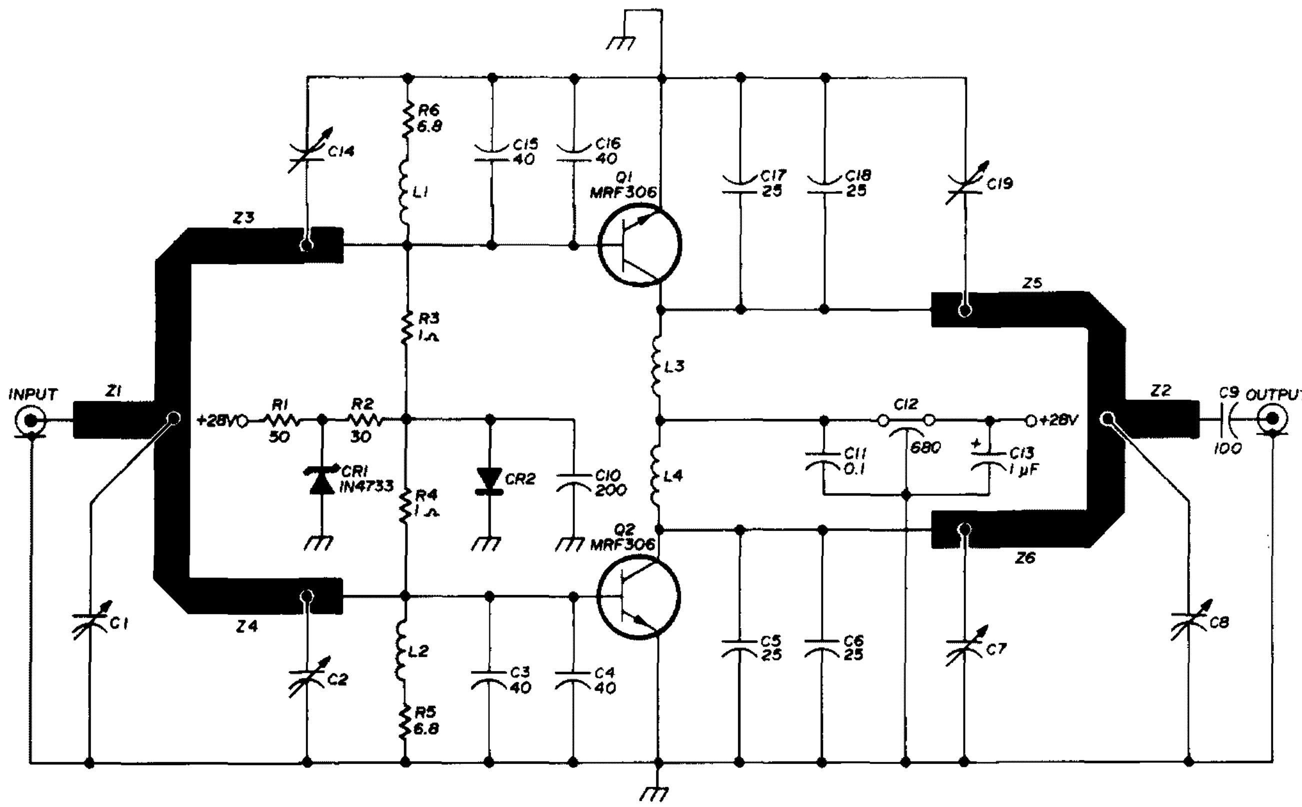


fig. 3. IMD performance of the 100-watt 432-MHz power amplifier vs output power (V_{CC} = 28 Vdc, quiescent collector current = 100 mA).



C1,C2, C14	20 pF mica trimmer (Arco 402)	C17,C18	25 pF metal-clad mica capacitor (Underwood)
C3,C4	40 pF metal-clad mica capacitor (Underwood)	CR2	1 amp silicon diode with 0.6 volt forward drop (Motorola MR501)
C5,C6	25 pF metal-clad mica capacitor (Underwood)	L1-L4	4 turns no. 22 (0.6mm) enameled wire, closewound on 1/8" (3mm) mandrel
C7,C8 C19	40 pF mica trimmer (Arco 403)	R1	50 ohm, 5 watt
C9	100 pF metal-clad mica capacitor (Underwood)	R2	30 ohm, 1 watt
C10	200 pF metal-clad mica capacitor (Underwood)	R3,R4	1 ohm wirewound, ¼ watt
C11	0.01 µF disc ceramic, 100 V	R5,R6	6.8 ohm, ¼ watt
C12	680 pF Allen Bradley feedthrough capacitor	Z1	Microstrip 0.11" (3mm) wide, 0.75" (19mm) long
C13	1 µF tantalum, 50 V	Z2	Microstrip 0.11" (3mm) wide, 1.3" (33mm) long
C15,C16	40 pF metal-clad mica capacitor (Underwood)	Z3,Z4	Microstrip 0.11" (3mm) wide, 2.3" (58.5mm) long
		Z5,Z6	Microstrip 0.16" (3mm) wide, 1.7" (43mm) long

fig. 4. Schematic diagram of the solid-state 100-watt linear amplifier for 432 MHz. This amplifier may also be used on fm or CW by replacing the class AB bias circuit with a ferrite bead and inductor as shown in fig. 5.

A 0.192 inch (5mm) hole is drilled through the strap for insertion of the capacitor and a small right-angle bend is placed in the strap so it can be soldered upright on the board.

The 1 µF tantalum capacitor, C13, is connected on the side which is electrically closest to the power supply. The 0.1 µF disc capacitor, C11, is connected on the other side of the feedthrough. The collector chokes L3 and L4 come straight off the feedthrough to a point on the collector leads as close as possible to the cap of the devices.

bias circuit

As you may or may not know, the h_{FE} (dc current gain) of transistors will vary from device to device but will always stay within a prescribed tolerance. As a result of some laborious experimentation, I have determined that the biasing scheme used in this amplifier will keep the MRF306s biased between 20 and 50 milliamps of quiescent collector current. This variation in quiescent current has little or no effect in IMD performance but does have a small effect on gain.

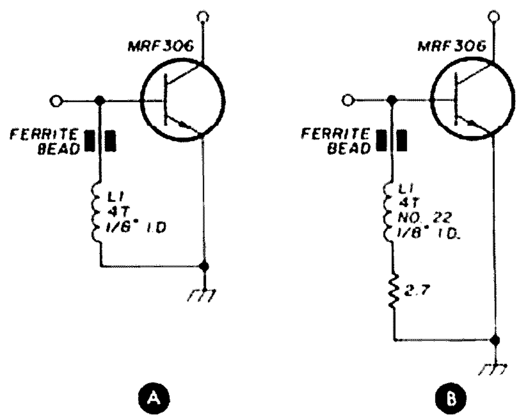


fig. 5. Power amplifier can be operated in class C for fm and CW by connecting a ferrite bead and inductor from the base of each device to ground as shown in (A). The devices can be driven further into class C by adding the 2.7 ohm series resistor as in (B). Inductor L1 is 4 turns no. 22, 1/8" (3mm) inside diameter.

The placement of the 1-ohm wirewound resistors, R3 and R4, is not as critical as that of the collector chokes; it is still advisable, however, to have the bias brought in as close to the transistor package as possible. Remember, too, that the 1-ohm wirewound resistor acts as an rf choke. Carbon varieties will not work in this application.

It is necessary to use approximately 100 nH of inductance in series with the 6.8 ohm resistors, R5 and R6, as they are carbon-composition resistors. The junction point of diode CR2, resistors R2, R3 and R4, and capacitor C10 use the isolated tab of the Unelco capacitor as a standoff. The anode lead of the zener diode, CR1, may be soldered directly to the circuit board. The cathode lead then serves as a tie point for

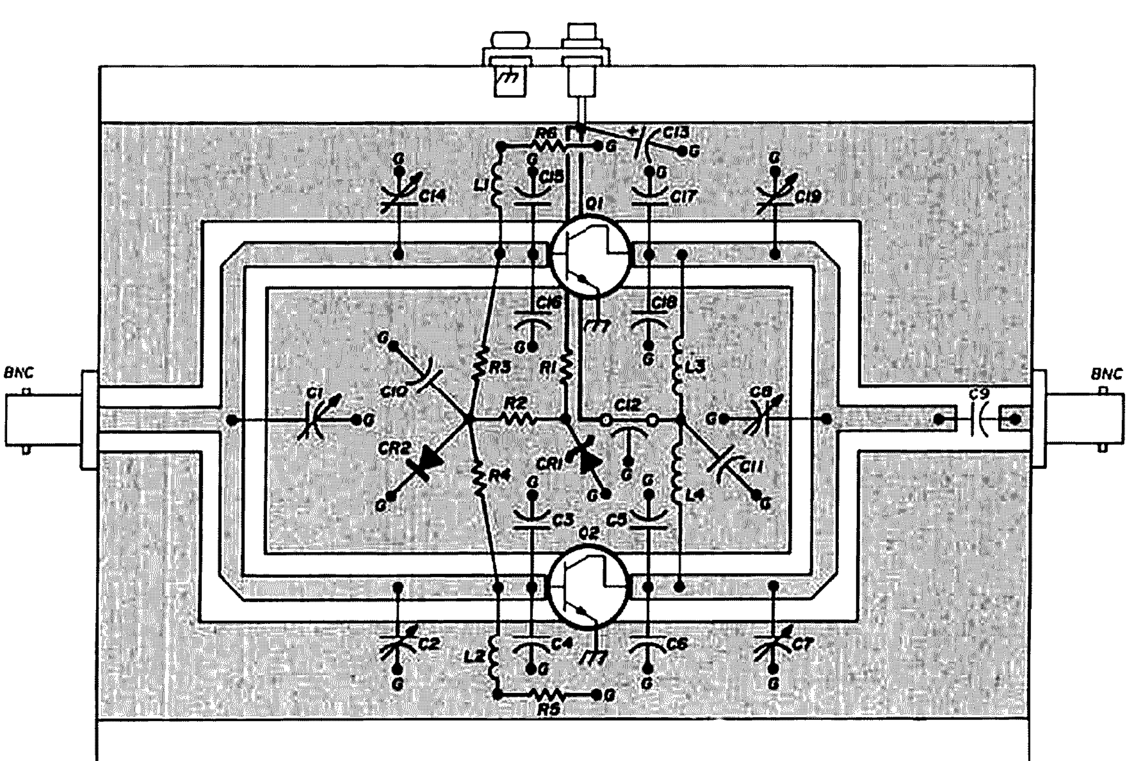


fig. 6. Component placement for the 432-MHz power amplifier. Full-sized printed-circuit board is shown in fig. 7.

resistors R1 and R2. The other end of resistor R1 is connected directly to +28 volts.

Finally, the 100 pF series capacitor in the output, C9, is mounted by placing it on its side on the microstrip with its isolated tab connected to the BNC output connector. This capacitor presents only a small impedance bump in the collector load circuit. Its main purpose is to prevent shorting the dc power supply to ground when using a dc grounded antenna such as a Yagi.

power supply

Let's digress for a moment and give consideration to a subject that is often neglected in articles such as these: A good power supply. Good IMD response is not only dependent upon bias, a good linear device, and proper collector loading, but also on a power supply that is capable of following the load changes presented by ssb operation. The power supply should have enough current capacity to handle peak current requirements and sufficient voltage regulation to prevent gross V_{cc} fluctuations during load and no-load conditions, as encountered during modulation. **Fig. 8** is a basic block diagram of a simple power supply that meets these requirements for this amplifier.

The transformer should provide from 26 to 35 volts with a current rating of 10 amps or greater. If you thumb quickly through practically any surplus parts catalogue you will find many suitable transformers at very reasonable prices. The bridge rectifier is anything capable of 10 amps. I recommend using the MDA962-1. A Motorola MPC1000 voltage regulator IC, followed by a large capacitor, will provide excellent regulation.

Tune-tup of the amplifier is not as tricky as it may seem, although it will require a little caution at first. It is advisable, if you have a means of varying the drive level, that you begin by applying only 3 to 5 watts to the input. It is very important that you get the

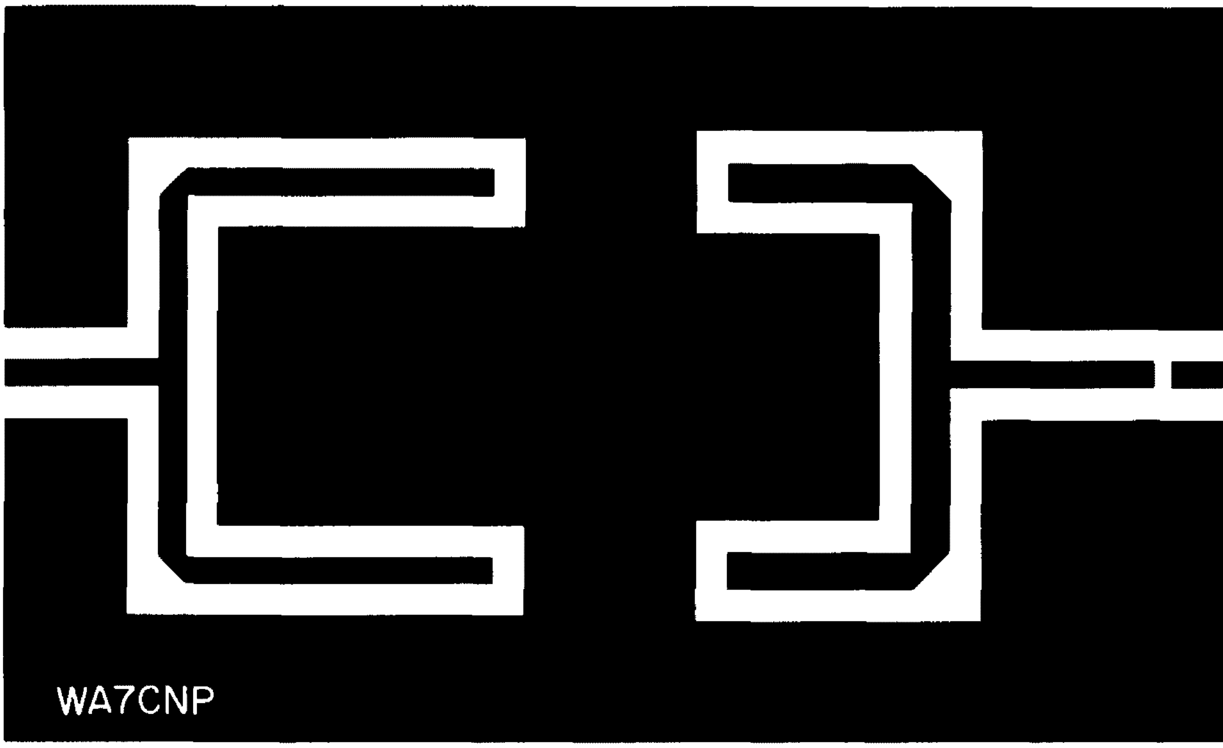


fig. 7. Full-sized printed-circuit layout for the 100-watt linear power amplifier. Component placement is shown in fig. 6.

collectors loaded to a "ball park" point before you start pouring on the coal.

When you first begin tune-up, adjust the input trimmers until you see a small deflection of output power. Then immediately adjust the output trimmers for peak output, ignoring the input match for the moment. The best way to adjust the output is to adjust C8 for a peak before adjusting C7 and C19. Capacitors C7 and C19 should be adjusted so that both Q1 and Q2 are equally sharing the load. Once you have a reasonably good collector load established, you can go ahead and do the same to the input match.

You will notice that, as output power increases, the output circuit will require small amounts of adjustment. This is because the required collector load impedance changes slightly as power increases. A small decrease in

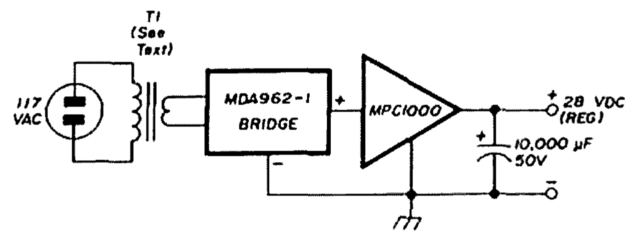
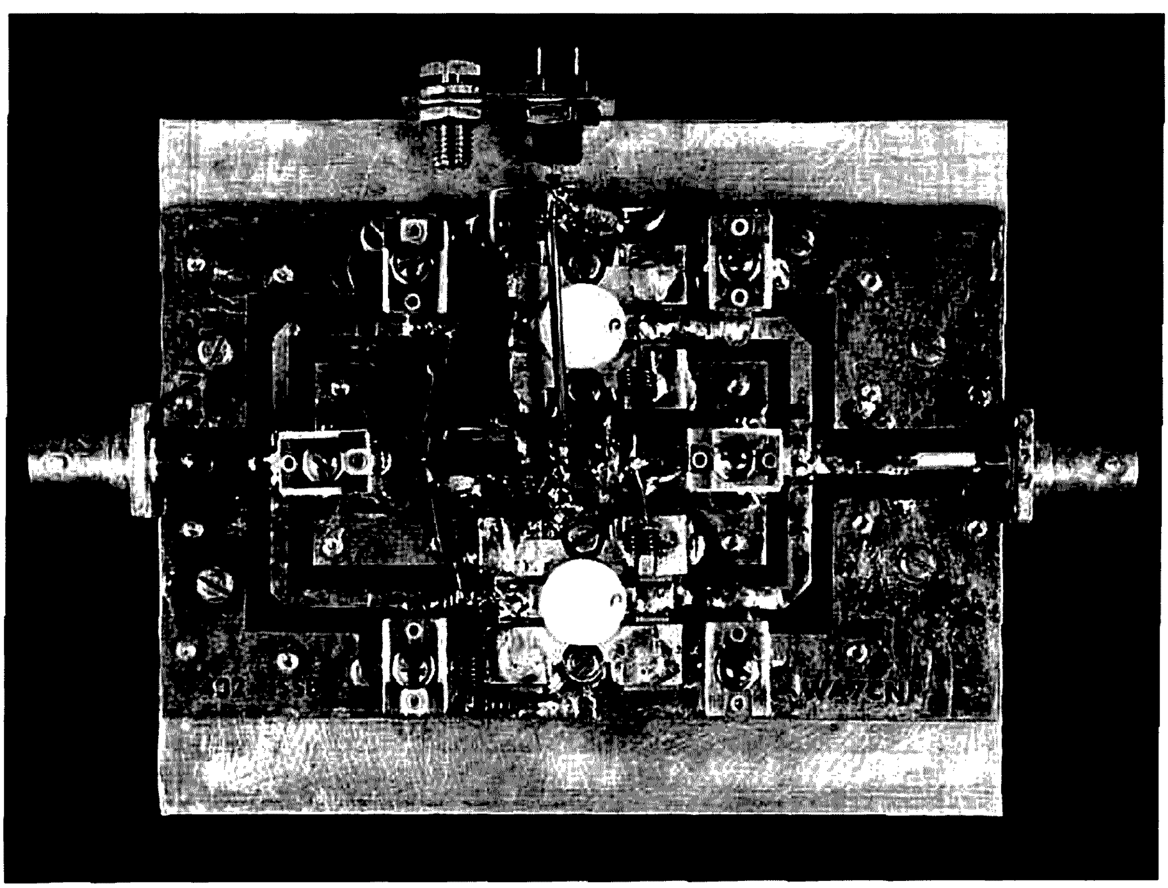


fig. 8. Basic block diagram of a regulated 28-volt supply for the 100-watt solid-state power amplifier. The Motorola MPC1000 is a positive voltage regulator designed to deliver up to 10 amps dc.

capacitance will probably be required on all the output trimmers as output power comes up. Alternately adjust the input and output circuits until you have reached your desired operating point (around 100 watts output).

The collector current should be between 6 and 8 amps depending upon the operating frequency. Capacitor C8



Top view of the solid-state 100-watt 432-MHz power amplifier, showing placement of the various components. Component designations can be easily correlated with the printed-circuit layout diagram in fig. 6. Input is at the left.

may be adjusted for slightly better efficiency but be careful not to drive the amplifier above 100 watts PEP output as severe flattopping will occur in the output waveform.

I hope you find this amplifier to be a useful tool in extending the capabilities of your 432 MHz station. There are many ways that this amplifier can be made to work even better, but it was my desire to make it as simple and inexpensive as possible without seriously degrading performance.

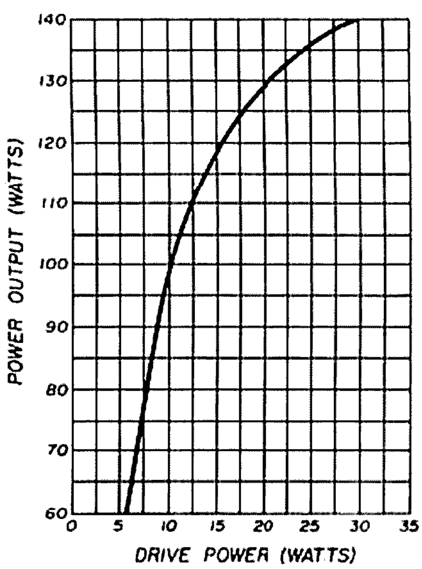
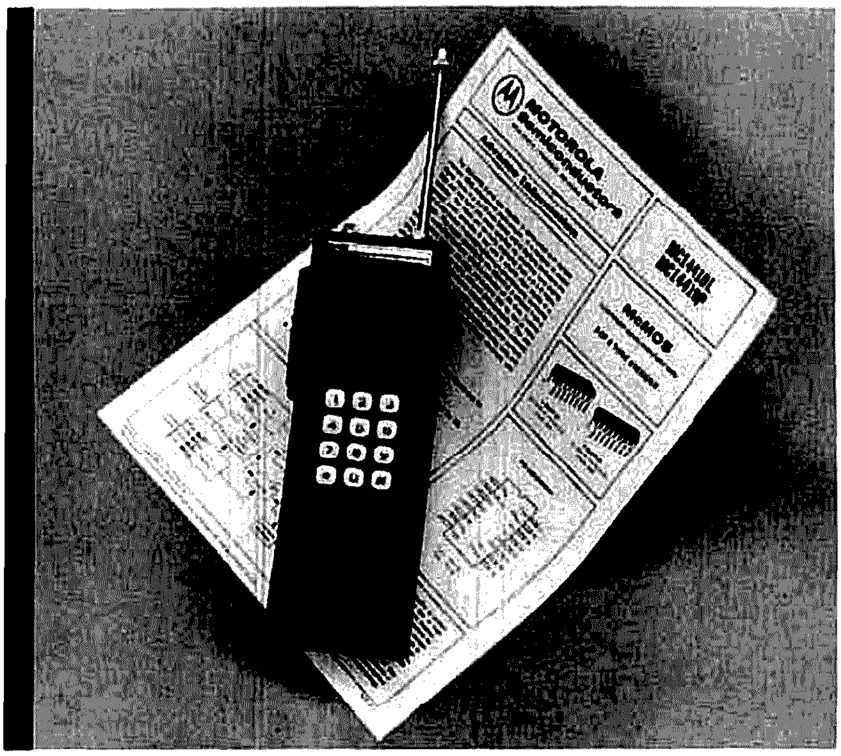


fig. 9. Plot of drive vs output power for the 100-watt 432-MHz power amplifier operating in class AB (V_{CC} = 28 Vdc). Note that the knee of the curve is just above 100 watts output; this corresponds to the rolloff in IMD performance shown in fig. 3.

A lot of the technology employed in solid-state power devices and amplifier design is new to amateur radio (and to industry as well) but as soon as technical publications and education programs can get geared up to this new technology, this type of work will become more and more commonplace. It should be the job of amateurs who are familiar with this technology to show the rest how it is done, and also to be the first ones to improve its implementation.

ham radio



hand-held touch-tone

New Motorola tone-encoder IC is the basis for this miniature Touch-Tone encoder for use in hand-held fm transceivers

Al Lowenstein, K7YAM, Motorola Semiconductor

The continuing increase in the popularity of autopatches, on both two meters and 450 MHz, coupled with the large number of hand-held fm transceivers now available, has generated the need for a truly compact tone encoder. With the introduction of a single-chip IC for this purpose, the handie-talkie Touch-Tone is now a practical achievement and for less than $35 you can enjoy all the benefits of your autopatch while operating portable with a handie-talkie.

The Motorola MC14410 is a complementary mos IC and is functionally a 2-of-8 tone encoder or Touch-Tone.[1] It is available in both plastic and ceramic packages, but for amateur purposes there is no reason to pay the premium for the ceramic package. Designed for 4.4 to 6 volts, it will withstand moder-

ate over-voltages. As it requires only 4 to 5 mA, it has little effect on the total current drawn by the radio and can be switched with the T-R key. The IC is complete within itself except for a 1-MHz crystal and the necessary components to limit voltage and match the output to the radio.

Unlike many other hybrids available today, the MC14410 is not susceptible to rf interference, so close proximity to the transmitter is no problem. The board shown here was designed to slide into the PL area of an Omni-housing Motorola HT220 radio. Fortunately, it is small enough to be tucked away in some corner of almost any fm transceiver. The crystal can be remoted on pigtails, and does not have to be mounted on the printed-circuit board, but can be located wherever space is available.

construction

Any good PC board material can be used for the circuit board. Since the solder pads are small, a board with good adhering foil is necessary; the phenolic materials seem to lack this attribute. Use a number-64 (0.9mm) drill and a high-speed drill motor to avoid tearing the pads. The IC should be soldered in last to avoid any possible static build-up on the MOS devices. Use a soldering *iron*, not a gun. Pins 1 and 7 are clipped off as they are not used.

Very flexible, fine gauge wire is advised for the board-pad leads; the leads should be approximately 2-1/2 inches (6.5cm) long. Solder all the wires to the board first; Teflon insulated wire is easier to work with because the insulation doesn't melt at soldering temperatures. The double pad on pin 11 of the IC is for the fourth column tone if you have a 4x4 keyboard – it was not needed here.

As shown on fig. 1, the 1-µF capacitor, C1, must be a nonpolarized unit. An alternative is to use two 1-µF electrolytics in parallel with reversed polarity. Since the output is approximately 25 kilohms to ground, the capacitor can be tied directly into the audio line.

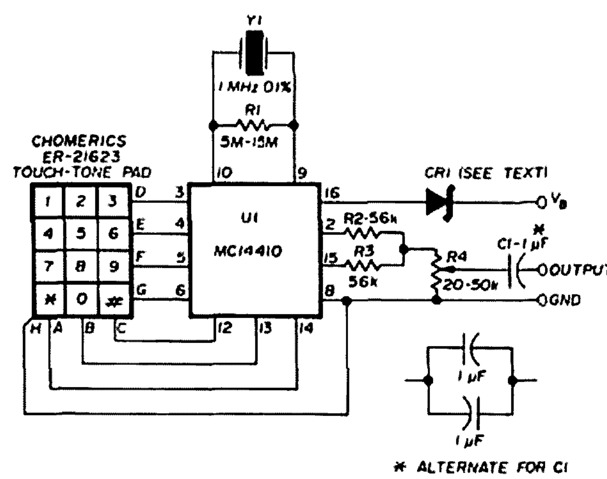


fig. 1. Schematic diagram of the miniature Touch-Tone encoder for the HT220 fm handie-talkie.

A zener diode, CR1, is used to limit the voltage to the IC at high battery conditions; it should be picked so that $V_{bat(h)} - V_z = 7$ volts. For the HT220, since the battery goes to approximately 16 volts at full charge, a 9-volt zener is used. Then, when the battery is low, there will still be 4.5 volts (13.5 – 9 = 4.5 Vdc) available for the encoder.

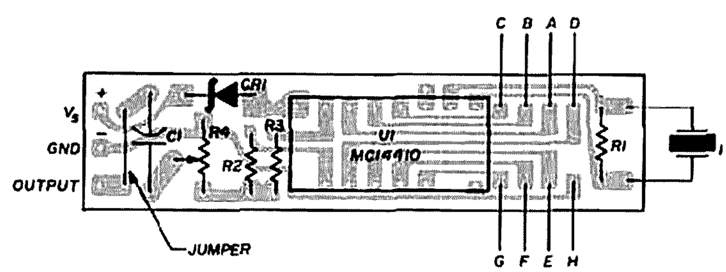


fig. 2. Component layout for the miniature tone encoder. Full-size circuit board layout is shown in fig. 3. Kits and components are available from Contact Labs, 35 W. Fairmont, Tempe, Arizona 85282.

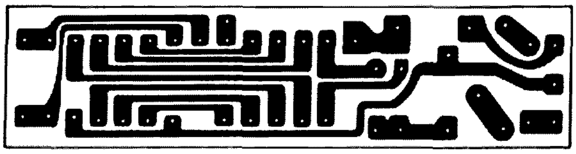
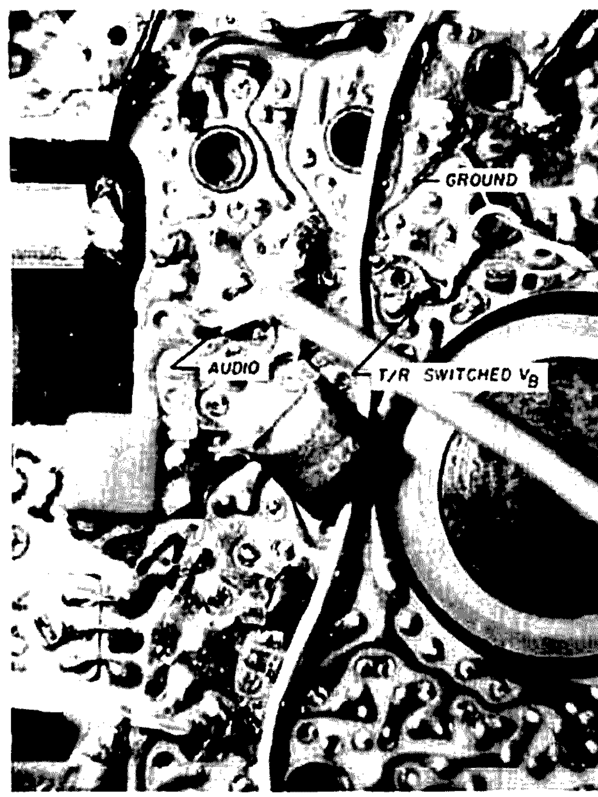


fig. 3. Full-size printed-circuit board for the miniature tone encoder.

The Chomerics ER-21623 pad pins must be clipped to approximately 1/8 inch (3mm) so they will clear the components of the HT220. Remove the fuse in the HT220 front cover from its clip while cutting the slots for the pins. An Xacto saw works well after pilot holes have been drilled; a Dremel Mototool could also be used. Extreme neatness isn't necessary as the Chomerics pad will completely cover the slots after it is epoxied into place.

The four plastic pins at the corners of the pad should be cut off and filed flush with the case. Fit thin pieces of cardboard around the pin areas to support the pad (see fig. 4). The pad is then epoxied to the front of the radio. Be sure to roughen the plastic of the radio with fine sandpaper to insure a good bond. Also, be careful to only epoxy the four edge rails of the pad. A little goes a long way, and epoxy can seep around the inside edges of the pad and make it inoperative.

After the epoxy has hardened, you are ready to put it all together. The pins of the pad should be tinned using a good clean hot iron. Do not overheat the pins as they will come free internally from the pad, resulting in erratic or no operation. The eight wires from the circuit board can then be soldered to the keyboard. The B+ supply, ground and audio are picked up as shown in the photograph. The difficult part is stuffing all those wires in. A gentle loop between the front cover and main circuit board will suffice, and there don't appear to be any detrimental effects from having the eight wires right next to the radio circuitry.



Interconnect points for the Touch-Tone encoder installed in a Motorola HT-220.

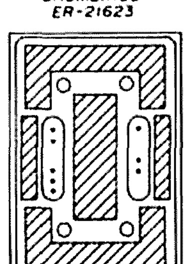


fig. 4. Cardboard inserts (hatched areas) are used to support the Chomerics Touch-Tone pad (see text).

The output potentiometer, R4, should be adjusted to produce a 4 to 5 kHz final frequency deviation when one key is depressed. This assumes, of course, that the radio's modulation causes a similar deviation during normal voice operation. Note that the microphone is live, so use precautions while dialing.

reference

1. Jon DeLaune, W7FBB, "Digital Touch-Tone Encoder for VHF FM," *ham radio*, April, 1975, page 28.

ham radio

how to use milliammeters

A discussion of meter shunts and other techniques to adapt surplus panel meters to the operating requirements of your own circuits

Guy Black, W4PSJ, 12317 Hanger Road, Fairfax, Virginia 22030

Quite often neither your junkbox nor the local surplus outlet has the meter you need for some new piece of equipment. It is very easy to substitute low-range microammeters or milliammeters where higher range milliammeters are wanted, or voltmeters in some cases. Meters have a generally-unknown internal resistance and a shunt resistor will divide the total circuit current flow between the meter and the shunt according to Ohm's law.

It is not necessary to determine the meter resistance to apply this ancient trick, although that is not particularly difficult, and meter shunts can be satisfactorily made with readily-available materials. For example, on a recent rig I shunted a 60 mA meter to read 600 mA, a 200 μA meter to read 20 mA and a 100 μA meter to read 10 mA. The basic test setup for current-meter shunting is a variable current source and a calibrated meter, either a small battery, with a series variable resistor, or an adjustable current-limiting power supply such as the Heath HP-28 which can be set at the desired full-scale current.

For milliammeters, shunt resistances are often very small: a 10 mA meter, for example, is likely to have about 2.5 ohms internal resistance and a 200 mA meter, about half an ohm. In the resistance ranges desired, copper wire makes quite satisfactory shunts.

meter shunt theory

For full-scale meter readings there will be an IR drop across the meter. If it is desired that the meter read full scale when the total circuit current is higher, the shunt must have whatever resistance

will pass all the additional current at the same IR drop. For example, a 10 mA meter with 2.5 ohms internal resistance will have an IR drop of (0.01 X 2.5 = 0.025 volt). If you wish the meter to read full scale when the circuit current is actually 100 mA, the shunt must pass 90 mA at 0.025 volt, which calls for a 0.278 ohm shunt (0.025V ÷ 0.09 mA = 0.278 ohm). A standard-value 0.27 ohm resistor will do the trick.

It is not difficult, however, to zero in on the right shunt value without knowing the internal resistance by taking a random length of small copper wire, placing it in series with the constant-current source, connecting the meter to one end and tapping it (with a needle-tipped probe) along the shunt wire until the meter reads full scale. Number-30 enamel wire, for example, has about 0.1 ohm per foot (3.28 ohm per meter) and the resistance and current-carrying capacity of other wire sizes appear in the ARRL Handbook. Thus, 2.8 feet (85.3cm) of number-30 wire wound in a small coil would make a suitable shunt for the previous example.

A lot of trial and error can be saved, however, by some simple calculations. Start by making a trial shunt, erring on the low resistance side if you have limited control over your constant-current source. Here is how it went shunting a 60 mA meter to 600 mA. With about 3 inches (76mm) of resistance wire, 60 mA in the circuit produced a meter reading of 15 mA, meaning that 45 mA was being passed by the shunt. As the IR drop across the meter and the parallel shunt is the same, two Ohm's law expressions can be stated in terms of three unknowns: the meter resistance, R_m, the test shunt resistance, R_s, and the common voltage across the meter and shunt, E,

$$\begin{aligned} E &= 0.045\, R_s \\ E &= 0.015\, R_m \\ R_s &= 0.333\, R_m \end{aligned}$$

When this meter is used in the 600 mA circuit the required current division between the meter and the shunt is 60 mA through the meter and 540 mA through the shunt. Using the same set of equations, the relationship of the meter resistance to the desired shunt resistance, R_s' is calculated as follows:

$$\begin{aligned} E &= 0.540\, R_s' \\ E &= 0.060\, R_m \\ R_s' &= 0.111\, R_m \end{aligned}$$

Since the meter resistance, R_m, is an unchanging value, the *desired* shunt resistance, R_s', compared to the *test* shunt resistance, R_s, is:

$$\frac{R'_s}{R_s} = \frac{0.111\, R_m}{0.333\, R_m} = 0.333$$

This is the same as saying that the wire in the desired shunt should be exactly one-third the length of the wire in the test shunt. The unknowns are still unknowns, but the desired result has been achieved.

When shunting sensitive milliammeters or microammeters, a low-range resistance decade is nearly essential because the resistance values are too high for lengths of copper wire. Alternatively, a small-value potentiometer can be used as a variable shunt, the value set for full-scale reading at the desired circuit current and then measured with an ohmmeter.

The approximate shunt values can be estimated by noting that 1 mA meters typically have about 30 ohms resistance; 500 μA instruments, about 90 ohms; 100 μA, 500 ohms; and 50 μA, about 2500 ohms. (If a decade resistance box is used, be sure it has shorting contacts or the full circuit current will instantaneously go through the movement during switching!) With this technique, the actual value of the desired shunt is determined and it is possible to select series or parallel combinations of stan-

dard-value resistors that will closely approximate the desired shunt value.

Note that decade resistance boxes can be constructed with only four resistances by using the Mallory type 154L switch, which automatically arranges a 1-ohm, two 2-ohm and a 5-ohm resistor so as to provide the entire decade range. In the one-tenth to one ohm range switch contact resistance is not likely to be serious. Such a decade will be very convenient if you plan to design shunts for very many meters.

There is less chance of blowing out meters accidentally in test set-ups if you wire the shunt in series with the current source and tap the meter around the shunt, rather than the other way round.

You will probably find that resistance wire is a bit hard to come by, and generally bare wire intended for heating elements is all that can be found. Standard Scientific Supply* carries Chrome-A resistance wire (80 percent nickel and 20 percent chromium) which has the best temperature/resistance relationship. Its resistance characteristics are listed in table 1.

critical damping resistance

If a meter is heavily shunted, the movement will be severely damped by self-induced current, and this may be undesirable in applications such as measuring the plate current of a linear, or VU instruments. The amount of damping can be reduced by putting a small resistor in series with the movement and shunting the combination of that resistor with the movement

An interesting experiment with a sensitive high-quality meter is to vary the shunt resistor with a potentiometer and observe how the needle oscillates before finally coming to rest. The largest resistance for which the needle approaches the final setting without overswing is called the critical damping condition. With more resistance there will be a slight overswing, and with less resistance the final setting will be approached more slowly.

In amateur applications there is no particular need to achieve critical damping, but it can easily be approximated by making the series combination of the shunt and resistance in series with the coil equal to the critical damping resistance.

table 1. Resistance characteristics of Chrome-A nickel-chromium resistance wire.

B&S gauge	diameter	resistance per foot (ohms)	resistance per meter (ohms)
18	0.0400" (1.02mm)	0.406	1.332
20	0.0320" (0.81mm)	0.635	2.083
22	0.0253" (0.64mm)	1.017	3.337
24	0.0201" (0.51mm)	1.610	5.282
26	0.0159" (0.40mm)	2.570	8.432
28	0.0126" (0.32mm)	4.100	13.451
30	0.0100" (0.25mm)	6.500	21.325

meter multipliers

Voltmeters can also be multiplied using additional series resistors, and low-current milliammeters or microammeters can be easily converted to voltmeter use. In this use, it is well to operate far below the wattage rating of the shunts because the temperature coefficient of standard carbon resistors is not the best. In addition, the maximum voltage across any one resistor should be kept below about 400 volts.

One of the best uses for shunts is for multi-circuit metering of equipment where different current or voltage ranges are desired for each circuit. In such configurations it is very useful to mount the shunts on the switch – generally a non-shorting type. You must watch the voltages on the various circuits, and on the meter as well. Current production panel meters from Weston and other firms built to government

*Standard Scientific Supply Corp., 808 Broadway, New York, New York 10003.

specifications are tested at 3000 volts (actually 2600 volts rms) between the movement and the case, but very old meters with metal zero-adjust screws should be used in high-voltage circuits only with great caution.

If a multiplier resistor is used to expand the range of a voltmeter which is switched off the multiplier, leaving the multiplier resistor floating, it will assume the full voltage potential which may be more than the switch insulation can stand. A better approach is to run the multiplier to ground, taking the voltage off a tap on the multiplier resistor. With this arrangement, the voltage across the switch contacts will be far less.

ayrton shunt

You can easily convert a sensitive meter into a multirange milliammeter with a few standard-value resistances and a switch using the configuration known as the Ayrton shunt shown in **fig. 1**. The values given will yield full-scale readings of 10, 100 and 1000 milliamperes within a few percent from a 1 mA meter with 50 ohms internal resistance. If your meter has less internal resistance, add enough resistance in series to bring it up to that value. A

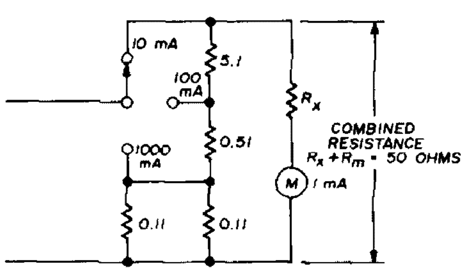
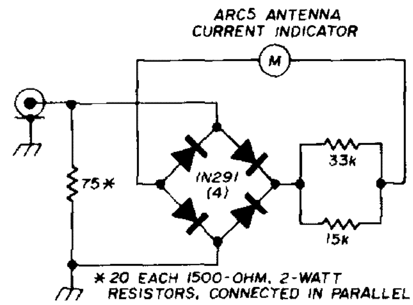


fig. 1. Using the simple Ayrton shunt circuit shown here, a 1 mA meter with 50 ohms resistance can be used to measure 10, 100, and 1000 mA.

rough way to measure internal meter resistance is to adjust a constant-current source through a resistor to bring the meter to full scale, and then add enough resistance in parallel to bring the meter to half scale. The *added* resistance then equals the meter resistance.

Meters are sometimes found with non-linear movements, and they can be particularly useful in some applications (and of course internal resistance cannot be found in the manner just described).



CALIBRATION	
SCALE	POWER
1	40 mW
2	150 mW
3	330 mW
4	610 mW
5	1.65 W
6	2.70 W
7	5.00 W
8	8.60 W
9	16.9 W
10	≈ 45W

fig. 2. This wide range rf wattmeter uses surplus non-linear "antenna current indicator" meter used in Command Set transmitters.

Many amateurs have such meters in their junkboxes which were salvaged from war surplus Command Sets, a 2½-inch (64mm) meter with a zero-to-ten indication labelled, "antenna current indicator." Half-scale of this meter is about 5 milliamperes, but it is very sensitive at the low range and very insensitive at the high range. Built into the rf wattmeter shown in **fig. 2**, a scale reading of 1 corresponds to 40 milliwatts on my instrument, 5 corresponds to 1.6 watts, and 9 corresponds to 17 watts. This wattmeter is for 75-ohm systems but a similar arrangement could be built for 50 ohms (I often use 75-ohm coax, as low-loss 75-ohm CATV cable is often available on the surplus market.

If you cannot find a non-linear meter you can achieve the same result by shunting any meter with diodes and add-

ing enough series resistance to the movement to bring the combined IR drop to about the contact potential of the diode. Table 2 gives the full-scale IR drop of a common (and typical) series of panel micro/milliammeters. Since silicon diodes start to conduct at about 0.6 volt, nonlinearity is achieved by adding enough series resistance to the movement at R_Y in fig. 3 so that the combined IR drop is in this general range.

If too large a resistance is added the meter will not reach full scale; if just enough resistance is added so that full scale can be achieved the reading will be very nonlinear, and by adding less the linearity is improved. In fact, merely adding a protective diode alone will produce some degree of non-linearity, although it is not likely to be noticeable except in microammeters.

Parenthetically, meters are protected only up to the point where the shunting diode blows out, and with high-current

table 2. Internal resistance of Simpson type 1212C panel milliammeters (typical of quality panel instruments).

full-scale reading	internal resistance (ohms)	full-scale IR drop (volts)
50 μA	5000	0.250
100 μA	2000	0.200
200 μA	1000	0.200
500 μA	200	0.100
1 mA	46	0.046
5 mA	23	0.115
25 mA	3	0.075
50 mA	1.5	0.075
100 mA	0.75	0.075
200 mA	0.6	0.120
1 amp	0.05	0.050

diodes available so inexpensively on the surplus markets, it makes sense to use a 10-ampere diode in any setup (such as an experimental bench supply) where short circuits are occasionally possible. These diodes need not be heatsinked.

Nonlinearity of the opposite sort can be achieved by putting zener diodes in series with meters, and some very odd curves of current versus needle deflection can be obtained by various combinations of zeners and shunt diodes. All of these effects, of course, ruin the original scale calibration.

Zener diodes in series with milliammeters result in expanded-range volt-

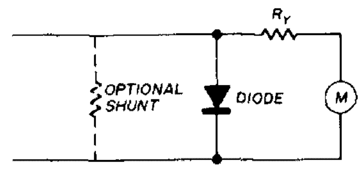


fig. 3. Simple circuit for non-linearizing meter movements uses semiconductor diode (see text).

meters. For example, a 150 volt, 10 watt zener such as the 1N3011, in series with a 1N4007 and 100-ohm resistor, spreads out 100 to 130 volts over the entire scale of a 50 mA meter. The amount of resistance determines the amount of compression. Lower voltage zeners – typically just above the target voltage – are very useful for expanded scale monitors of battery packs and transmitter filament voltages. The additional series diode is only necessary when monitoring ac.

summary

Liberal use of meters can be one of the design features where the home builder can outshine the commercial source. Good quality panel meters are still in abundant supply in surplus and at flea markets, often at very low prices. Changing scales and ranges is not very difficult for many types, though it is not possible to open up the newer hermetically sealed-instruments. The more odd-ball the scale, of course, the lower the selling price.

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1. Ernest Frank, *Electrical Measurement Analysis*, McGraw-Hill, New York, 1959.

ham radio

magnet-mount mobile antenna

This simple, magnet-mount vhf antenna is useful for temporary mobile operation where a permanently mounted antenna is not available

George Allen, W1HCI, 80 Farmstead Lane, Windsor, Connecticut 06095

A magnet-mount antenna is an indispensible item for anyone currently working vhf fm on 144, 220, or 450 MHz. When used in conjunction with a cigarette lighter adapter to provide 12 Vdc for the fm rig, it permits mobile operation from almost any vehicle at a moment's notice. Thus, from an emergency standpoint, you are not restricted to a vehicle already equipped with a vhf antenna. In addition, the magnet mount is useful for those who do not wish to cut a hole in their car to operate mobile.

Unfortunately, commercially available magnet-mount antennas are expensive, and up to now it has been difficult to obtain magnets with the right surface area, the right holding power, and the right price. This article describes a very simple quarter-wave magnet-mount antenna which can be built in about a half hour for any of the three vhf bands mentioned. The total cost of the antenna, not including coaxial cable, is less than $6.00. All parts are readily available, and the magnets can be obtained locally.*

The magnet-mount antenna is really too simple for words; **fig. 1** shows a drawing of the antenna. It simply consists of a good quality magnet, a 1½-inch (38mm) long 6-32 or 8-32 (M3.5 or M4) flat-head screw, three nuts, a lock washer, coax connectors, and a cable clamp.

construction

To build the antenna, connect a female uhf chassis connector to the magnet as shown in **fig. 1**. Run the coax under the cable clamp (or tape the coax to the screw if you don't have the clamp), and connect the inner conductor to the center of the coax connector and the shield to the flange. Take a piece of brazing rod of the proper length (see **fig. 1**), and solder it to the male uhf connector. Use some silicone

*If you can't find a local source, the 1½ inch diameter magnets are available for $3.95 plus postage from George Allen Engineering, 80 Farmstead Lane, Windsor, Connecticut 06095. Connecticut residents add 6% sales tax.

bathtub caulking such as that made by GE to fill up the male connector to provide a good seal. The brazing rod can be obtained from any welding supply house.

In regard to the magnet, use round button or shallow pot Alnico magnets or equivalent of 1¼ or 1½ inches (32 to 48mm) diameter. Magnets of this size provide enough holding power, while smaller magnets won't hold well and will blow off your car at high speeds. Note that when your antenna is not being used, make sure that the "keeper," the small metal piece provided with the magnet, is placed across the poles of the magnet. The use of the keeper prevents loss of magnet power.

tuning

Although this simple magnet-mount antenna will give you good results, it is not quite as efficient as a quarter-wave antenna permanently mounted on your car. In most cases, however, you probably couldn't see the difference between the magnet-mount antenna and one that is permanently mounted. In regard to feedline matching, it is difficult with this type of antenna to get the swr down to 1:1. However, this really doesn't matter since your feedline is very short and the overall losses will be small.

To tune the antenna, place the antenna on the center of the car roof or trunk and connect an swr bridge in the line between the transmitter and the antenna. Start snipping off pieces of brazing rod by 1/8 inch (3mm) at a time until the swr falls between 1.5:1 and 2:1. At this point the antenna is tuned, and the only remaining thing to do is to put some silicone seal or tape on the end of the antenna to protect against injury. Note that the swr may vary when the antenna is used with different cars.

APPROXIMATE LENGTH FOR QUARTER-WAVE 3/32" (2.5 mm) DIA. WHIP

MHz	IN.	cm
144	22	55.9
220	15	38.1
450	8	20.3

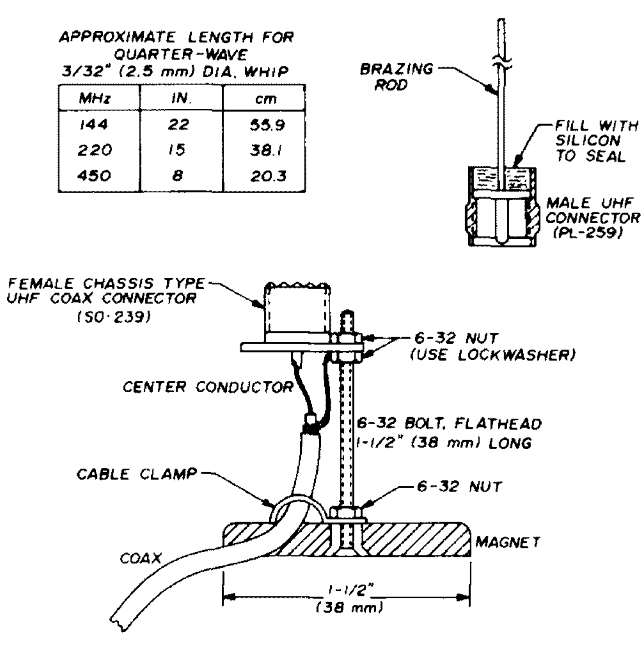

fig. 1. Construction details of the simple magnet-mount mobile antenna. Dimensions are given for quarter-wave operation on 144, 220 and 450 MHz.

performance

I have been doing quite a bit of traveling in the last six months and have been taking this antenna with me. It is convenient to use, works well, and has yet to come off a car roof at speeds up to 80 mph. I have been using the antenna on all sorts of rental cars and have had no problems except where the car has a vinyl roof. In those cases I place the antenna on the trunk. The magnet won't hold well through the thickness of the vinyl.

In practice, I place the antenna on the center of the roof or trunk and run the coax through a window, leaving the window slightly open so as not to damage the cable. I connect the coax to my rig and I'm on the air! It does not appear to be necessary to protect the car paint from the magnet. I have been using these mounts for four years and have yet to scratch the paint with a bare magnet.

I hope that this short article gives you the incentive to build this simple antenna; it's handy either for daily use or to keep in the closet for use in that emergency when you can't use your own car.

ham radio

300-Hz crystal filter for Collins receivers

Jim Fisk, W1DTY, Ham Radio Magazine, Greenville, New Hampshire 03048

Recently a number of 455-kHz Collins crystal filters with a 6-dB bandwidth of 300 Hz have become available on the surplus market.* These filters, which were built to military specifications, have a 60-dB bandwidth of 1200 Hz and maximum insertion loss of 5 dB, are designed for a source and load impedance of 2000 ohms, and do not require any resonating capacitors† (mechanical filters for Collins receivers are designed for terminations of 50 kilohms or greater and require external capacitance to resonate the transducer coils). If the crystal filter is not terminated with 2000 ohms, passband ripple will be on the order of 6 dB or more and spurious response will seriously degrade skirt selectivity.

Since the filters in Collins receivers are isolated by dc blocking capacitors, the required terminations for the crystal filter are most easily provided by simply connecting 2200-ohm resistors across the input and output terminals. Be sure to remove the 100 pF resonating capacitors from the circuit, however, as they will cause excessive passband ripple and unwanted spurious response. When terminated with 2200-ohm resistors passband ripple is nil and the skirts roll off smoothly to 80 dB or more.

Unfortunately, however, this simple resistive loading results in a serious impedance mismatch which manifests itself as 10 to 12 dB of additional circuit loss. Increasing the terminating resistors to 3900 ohms will reduce the loss about 3 dB, but passband ripple starts to suffer. A better solution is to drive the crystal filter with the simple emitter follower circuit shown in **fig. 1**. This circuit, which requires only 10 mA of current, reduces circuit loss to 3 dB or less and provides the filter with the required source impedance.

The emitter follower can be built on a small section of perforated circuit board which is supported by the input and output wiring. Power is derived from the screen circuit of the mixer tube. Make sure that the emitter follower is properly isolated with dc blocking capacitors as any dc voltage on the filter transducers will damage them (the filter switch has shorting contacts, so any voltage on the switch may damage adjacent filters as well). However, if you follow the circuit shown in **fig. 1**, which is completely isolated, you will have no difficulties.

Installation of the filter in 75S3B and later model S-line receivers requires only a length of number-20 (0.8mm) tinned bus wire, a lockwasher, and a 4-40 nut. The filter is installed below the chassis, on one side of the filter shield compartment, as shown in **fig. 2**

*The Collins 300-Hz crystal filters, X455KF300, with data sheets, are available from Gary Fertik, W1EBC, 40 Pilgrim Trail, Woodbury, Connecticut 06798. Price is $49.95, postpaid.

†XF455KF300 filters, Collins part number 526-7073-010. Other Collins crystal filters with the same generic nomenclature but with different part numbers are designed for 20k terminations; some require external capacitors. Although the filter described here is the most common, check the data sheet which comes with your filter.

(installation suggested by WA8OBG). Three holes are required: two for the electrical terminals and one for the mounting screw. Since these holes are below the chassis the filter installation does not deface the receiver.* The filter is symmetrical so either end may be used for the input or output.

Collins 75A4

Owners of Collins 75A4 receivers should be particularly interested in the 300-Hz crystal filter as the narrowest bandwidth filter designed specifically for this receiver has a 3-dB bandwidth of 500 Hz, and these filters are very difficult to find on the open market. Although there are two methods of installing the 300-Hz crystal filter in the 75A4, the installation shown in the photograph is recommended because it provides maximum isolation between the input and output (this same method is also recommended for Collins type-FA mechanical filters).

Turn the 75A4 upside down on your bench (front panel forward) and remove the bottom cover. The three filter sockets are in the front right-hand corner next to the selectivity switch. The crystal filter is installed in the shield which crosses the three filter sockets. Two 1-inch (25mm) deep slots must be cut in the shield as shown in fig. 3. Use *sharp* tin snips and place rags underneath the work area on both sides of the shield to catch any debris. After cutting the slots, bend the tab toward the rear of the receiver so it forms a 90-degree angle with the shield. A hole for the filter mounting screw is drilled in the center of the tab, ¼ inch (6mm)

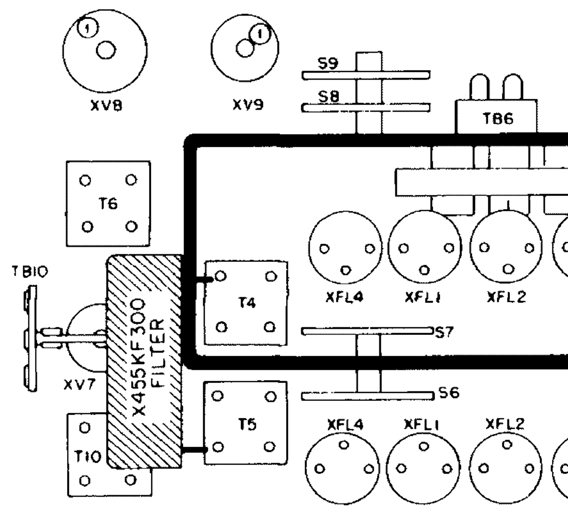

fig. 2. Installation of the Collins X455KF300 crystal filter in S-line receivers.

from the end (the threaded stud on the other end of the filter is not used).

After the tab is finished, temporarily set the filter in place to check for clearance between the top of the filter and the bottom cover of the receiver (if you follow the dimensions shown in **fig. 3**, the top of the filter should be approximately flush with the top of the shield).

Locate the input and output wires to filter socket A (underneath switch S2) and their respective connection points on the switch wafers (note that one of the wires is grounded). Remove the two 100-pF resonating capacitors. Install the emitter follower circuit shown in **fig. 1** between switch S2 and the crystal filter. The emitter follower common is con-

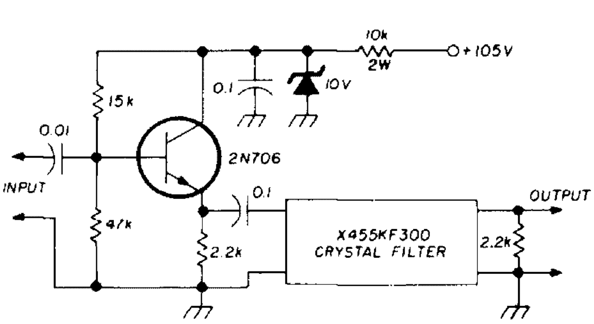

fig. 1. Emitter-follower circuit which provides correct source impedance for the crystal filter and minimizes circuit loss. Most general-purpose npn transistors may be substituted in the circuit.

*In *some* Collins 75S3B receivers there is sufficient clearance under the filter shield on the top of the chassis that the crystal filter can be installed in the existing crystal-filter sockets. Although the terminals of the X455KF300 will not fit the sockets, short lengths of no. 20 (0.8mm) wire can be soldered to the filter terminals and plugged into the sockets.

nected to the grounded terminal on the rear wafer of S2; the input coupling capacitor, C1, is connected to the other switch terminal which goes to filter socket A. Install two short lengths of number 20 (0.8mm) bus wire to each of the connection points on the front wafer of switch S2. (If you don't want to include the emitter follower, connect bus wires to the rear wafer as well.)

Connect a 2200-ohm resistor across the output terminals of the crystal filter and install the filter on the mounting tab using a lockwasher and 4-40 nut. Wire in the emitter follower and solder the two bus wires to the output terminals. Total installation time should be two hours or less.

Since you have the bottom of the receiver open, this is a good time to apply some contact cleaner (such as GC Electronics *Tunerlube*) to each of the switch contacts. It's also a good idea to dab some silicone grease on the switch detent mechanisms. If your 75A4 is like most, the only lubrication the receiver has ever seen was applied at the factory, and that's pretty well dried up. A little switch care now may save an expensive replacement problem later.

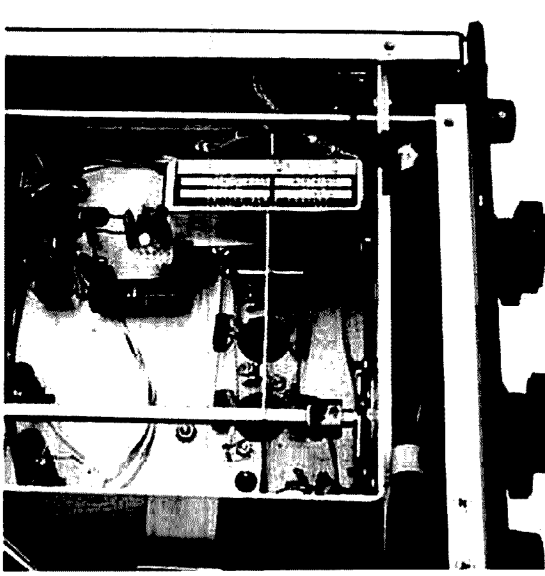

fig. 4. Crystal filter installation in the Collins 75A4 receiver. The emitter follower is on the small circuit board to the left.

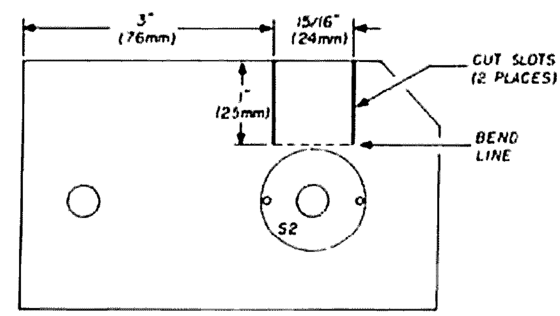

fig. 3. Collins 75A4 filter shield, showing modifications required for installing the crystal filter. Completed installation is shown in the photograph.

An alternative method is to mount the crystal filter on a small L-shaped bracket which is attached to the chassis with the screw which holds the left-hand end of the filter shield (next to the i-f gain control). In this case the filter connections are made to filter socket C. However, this method is not recommended because the connecting wires are quite long and lowered input-output isolation degrades the high skirt selectivity of which the filter is capable.

operation

If another mechanical filter is installed in filter socket A, it must be moved to socket B or C. Operation of the 75A4 with the sharp 300-Hz filter requires some practice to gain full advantage of its high skirt selectivity. The setting of the *passband tuning* control is quite critical and for best results should be set so that CW signals peak at a pitch of about 700 Hz. When you tune in a signal with a broader filter, set the main tuning for a 700-Hz note before switching the sharper 300-Hz filter into operation. If the signal is tuned for a higher or lower note (assuming the passband tuner is set for 700 Hz), the receiver must be retuned slightly to find the signal. With a little practice, you'll find that the narrow bandwidth and high skirt selectivity of this filter do an excellent job of cutting interference or digging into the noise for weak signals.

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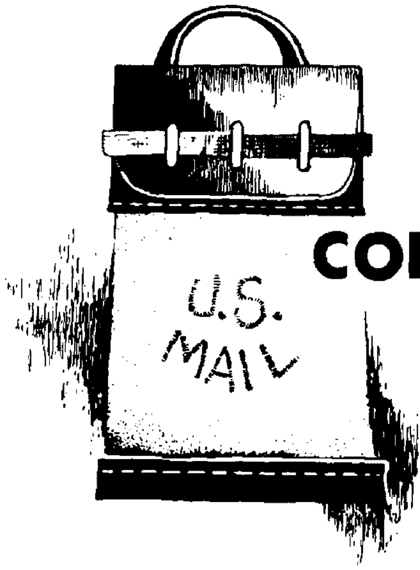


comments

touch-tone decoder circuit boards

Dear HR:

Due to the rather large response to my article on Touch-Tone decoders which appeared in the December, 1974, issue of *ham radio*, I have decided to offer printed-circuit boards for sale to those individuals who might want them. I am also putting together some kits for those builders who have a hard time finding all the necessary components. All the parts and materials are top quality, commercial grade and ICs will be pre-tested to ensure performance. A set of boards is $12.95, and a kit, including boards and toroids, is $37.50.

John F. Connors, W6AYZ
Electromedics
3295 Brookdale Drive
Santa Clara, California 95051

ssb transceiver

Dear HR:

A small number of the transceivers built from the ssb transceiver article in the August, 1974, issue of *ham radio* suffer from apparent agc instability. The symptoms are generally motor-boating at certain signal levels.

The problem is not, in fact, due to the agc but to instability caused by i-f feedback through the unused transmitter section of the circuit. It may easily be cured by connecting a single 0.1-μF capacitor with low rf resistance between the transmitter section power supply line and ground (as near as possible to the SL610C amplifier). Installing this capacitor does not remove the necessity of grounding the transmitter power line during reception and vice versa.

I apologize to anyone who has been inconvenienced by this fault, but the majority of these transceivers are not affected and the problem has only recently been brought to my attention.

Brian D. Comer, G3ZVC
Plessey Semiconductors

quadrature-phased local oscillator

Dear HR:

I have been following the articles concerning direct-conversion receivers and find the communication between Madey and Shubert regarding the Phase II receiver in the June, 1974 *comments*

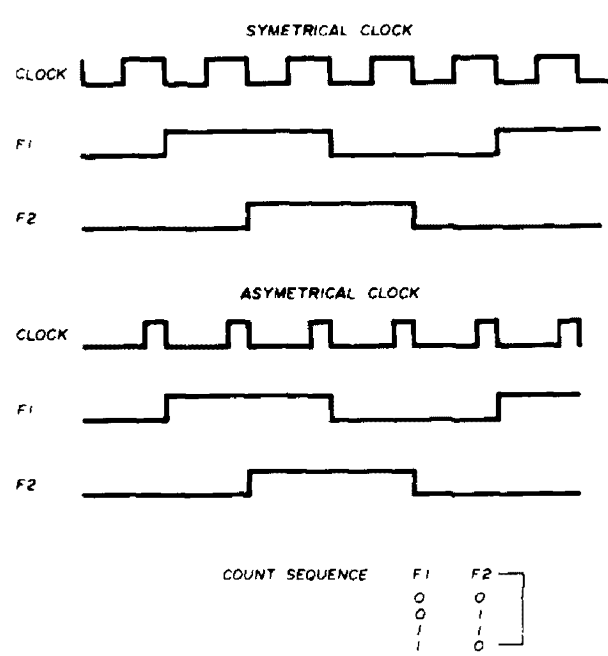


fig. 1. Waveform of the quadrature-phased local oscillator, showing independence of local-oscillator quadrature on clock-period asymmetry. Asymmetry of negative-going transitions would imply gross frequency instability.

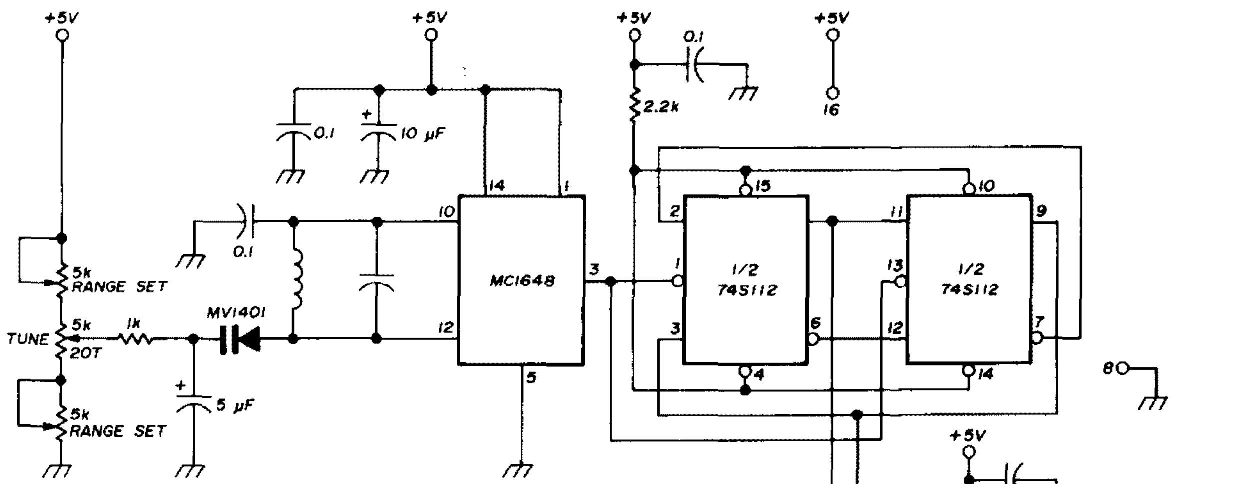
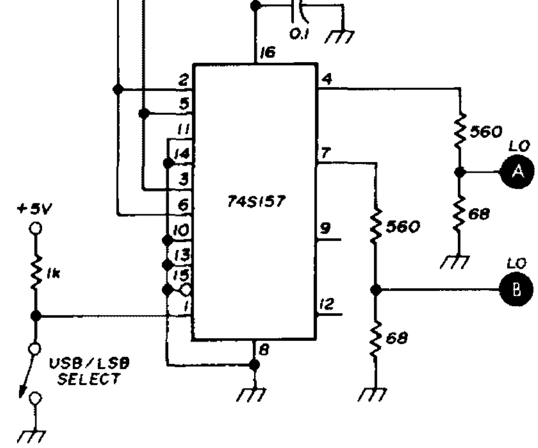



fig. 2. Schematic of a direct-phasing quadrature local oscillator designed for a direct-conversion 80-meter receiver. Timing sequence is shown in fig. 2.

section most interesting. The diagram of a quadrature-phased local oscillator I have been working on for use in a 3.5- to 4-MHz receiver is shown in **fig. 2.** As the timing diagram **(fig. 1)** demonstrates, the design is free of asymmetry errors since it is responsive to the negative-going transition of the clock waveform, and the clock may exhibit any periodicity it wishes, within device limitations. The circuit was intended for use with MC1496-type product detectors and has provision for switching the phase of the local oscillator to effect sideband reversal rather than performing the task at audio and having to accept a compromise in unwanted sideband rejection.

Douglas K. Beck, K6ZX
Sunnyvale, California 94086

Collins 75A4 mods

Dear HR:

Recently, when the avc failed in my Collins 75A4, the usual changing of tubes had no effect. Actually, a small amount of avc action remained – the S-meter needle rose slightly off zero with very strong signals. The trouble proved to be R86, a 39k resistor. Both R86 and R87 had suffered severe overheating – R86 had changed in value from 39k to approximately 3k, causing overheating of R87 and eventual failure of the avc.

My 75A4 manual lists R86 as a half-watt resistor. However, a friend has a later 75A4 manual, and it shows a rating of one watt (the serial number of my receiver is in the 2500 series). If you experience avc failure in your 75A4, first check R86 and, if you're working on the receiver anyway, make sure that R86 is a one-watt resistor.

Incidentally, I cannot recommend too highly the 75A4 mixer mods described by W6ZO in *ham notebook*. I installed them over a year ago and have been extremely pleased with the results. I also changed the first rf amplifier from a 6DC6 to a 6GM6 as recommended by W2VCZ, and recommend that, too, as it increases gain and sensitivity. However, I would not plug in the 6GM6 without first installing the W6ZO mixer modifications.

Bob Locher, W9KNI
Deerfield, Illinois 60015

Heath HM-2102 wattmeter mods

Dear HR:

With reference to the *item on the* Heath HM-2102 wattmeter by VE6RF,* the following additional information might also be of some interest. It was interesting to read how one amateur solved the problem of calibrating his Heath HM-2102 below an swr of 1.5:1. In my case the problem was solved in a slightly different manner.

In my initial calibration, using a Bird Termaline wattmeter, the minimum swr null was about 1.3:1. This was within the specified limits called for in the Heath instruction manual and, for all practical purposes, should have sufficed. However, in actual tests erroneous readings were obtained, often below the 1.3:1 reference level. An inspection of the schematic shows that C3 in series with C4 (the trimmer) will produce a capacitance of 1.99 to 5.56 pF. The total capacitance would then have to be changed to either a value lower than 1.99 pF, or higher than 5.56 pF. A 2 pF capacitor placed across C3 showed that the null could not be brought down to less than 2:1. Obviously the total capacitance of C3 and C4 had to be decreased instead of increased.

Cutting the long lead from C3 to the circuit board and inserting several values of capacitance in series with C3 showed an immediate improvement in the null. In my case a 10 pF capacitor brought the null down to a 1:1 swr. The total range of all three capacitors in series is now 1.66 to 3.57 pF. Replacing C3 with a capacitor of about 4 pF would probably have produced the same results. Since the capacitors in this circuit form an ac voltage-divider network, changes in this circuit will also affect the wattmeter reading so the wattmeter will also have to be recalibrated.

*R. Fransen, VE6RF, "Better Balancing of the Heath HM-2102 Wattmeter," *ham radio*, January, 1975, page 56.

In my case no thought was given to changing capacitor C16 since this appears to be a bypass for frequencies outside the desired range of 50 to 160 MHz. Changes here could influence the sensitivity of the bridge and may even bypass energy at the wanted frequencies.

In addition to the above, I found two other slight modifications to be useful. The first concerns wattmeter calibration. According to the instruction manual, power calibration is performed with the wattmeter in the 25-watt range; no provision is made for calibrating in the 250 watt range. However, after calibrating the wattmeter on the low range using 20 watts of power, switching to the high range showed a meter reading of about 16 watts. An examination of the schematic shows that if R8, a 68k resistor, is replaced by a fixed resistor and potentiometer in series, a second calibration can be made on the 250-watt range which is quite accurate. In my case R8 was replaced by a 51k, 1/4-watt fixed resistor in series with a small 25k potentiometer.

The last modification was for convenience more than anything else. In order to locate the remote chassis closer to the coax feedline, the short piece of five-conductor cable supplied with the kit was replaced with similar cable about 6-feet (1.8m) long. One end of the cable was connected inside the cabinet in accordance with the instructions, but the other end was terminated in a small five-pin plug instead of being connected directly to the remote chassis. A matching five-pin socket was mounted on the chassis off to one side of coax connector A. Short leads were then run from the circuit board to the five-pin socket. To prevent rf from reaching the indicator unit through the cable, the ferrite beads supplied with the kit were mounted at the socket instead of on the circuit board. Now the indicator unit and remote chassis can be easily separated.

B. T. Ring, K3VNR
Riverdale, Maryland

the ham notebook

non-synchronous impedance transformer

In matching one impedance level to another, as in antenna work, the usual device is a quarter-wavelength transformer section whose characteristic impedance is the geometric mean of the two impedances to be matched (fig. 1). When it is necessary to match one transmission line to another, and particularly in coaxial line applications, obtaining a length of line of the proper intermediate characteristic impedance may present a major problem.

The non-synchronous transformer shown in fig. 2 offers a way out of this dilemma. It is composed of two abbreviated lengths of transmission line of the same characteristic impedances as the impedances being matched. The procurement problem is therefore greatly simplified. As indicated in fig. 2, the transformer section length can be seen to vary according to the impedance ratio. For example, if a 2:1 impedance transformation ratio is required, each section of the transformer would be 28° 8′, making the total length about 56¼ degrees, or in terms of wavelength, 0.156 λ.

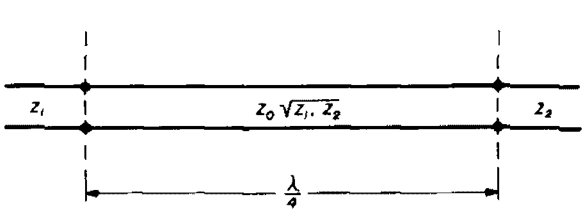


fig. 1. Characteristic impedance of the quarter-wavelength transformer is equal to the geometric mean of the impedances being matched.

It must be pointed out that the non-synchronous transformer is not always interchangeable with the quarter-wavelength version. The quarter-wavelength transformer will match an infinite number of impedance pairs, as long as their geometric mean is equal to the characteristic impedance of the transformer. The non-synchronous transformer, on the other hand, will match only the impedance pair for which it was designed. Within these limits, however, the non-synchronous transformer compares very favorably in

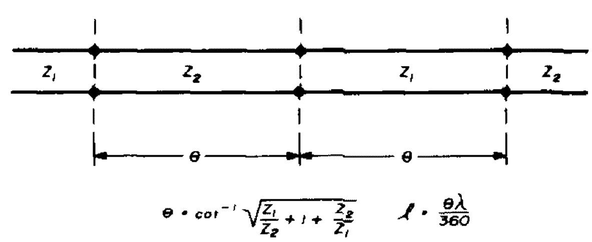


fig. 2. The non-synchronous transformer consists of two lengths of transmission line of the same characteristic impedance as the impedances being matched, with length varying according to the impedance ratio.

bandwidth as well as impedance-matching characteristics, and should find ready application in coaxial impedance-matching networks.

Henry Keen, W5TRS

drilling aluminum

The following hint for working with aluminum, which may not be common knowledge, was given to me by ham friends in Portugal: apply a drop or two of alcohol when drilling aluminum. It not only makes the work easier but results in a much cleaner cut.

Ralph Cabanillas, Jr., W6IL

metric conversions for screw and wire sizes

Here's a conversion chart you can use for plugging in metric values for machine screws and wire sizes. *Ham radio* articles have been including metric equivalents for dimensions of physical quantities such as area, length, mass, temperature, and volume. We have wanted to include metric equivalents for machine screws and wire but have only recently been able to obtain equivalent data for this hardware from the International Standards Organization (ISO). The ISO standard has not yet been adopted by all countries, but these tables will at least give overseas readers an idea of what size the author specified in the nearest metric standard.

Jim Fisk, W1DTY

table 1. Thread conversion, American to metric.

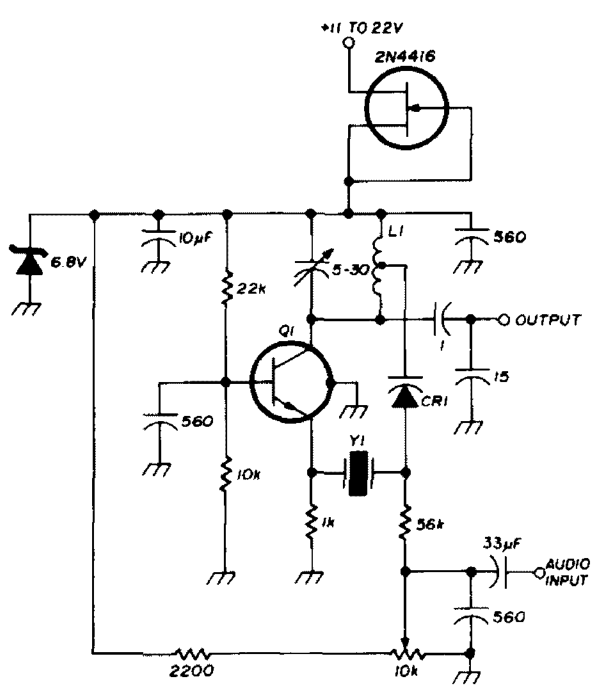
American standard	nearest metric standard
0-80	M1.6
1-64	M1.8
2-56	M2
3-48 or 3-56	M2.5
4-40	M3
6-32	M3.5
8-32	M4
10-24 or 10-32	M5
12-24 or 12-28	M6
1/4-20	M7
3/8-16	M10
7/16-14	M12
1/2-12	M12

table 2. Wire-size conversion, Amercian (AWG) to metric.

American standard (AWG)	nearest metric standard (mm)
2	6.5
4	5.2
6	4.1
8	3.3
10	2.6
12	2.1
14	1.6
16	1.3
18	1.0
20	0.8
22	0.6
24	0.5
26	0.4
28	0.3
30	0.25
32	0.2

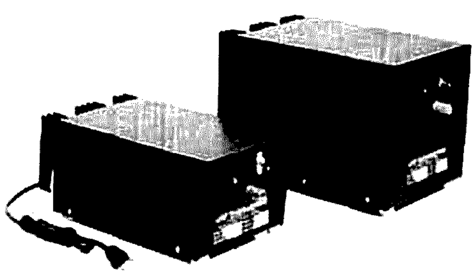
short circuit

In DJ2LR's excellent article on crystal oscillators in the June, 1975, issue of *ham radio*, there was an error in **fig. 5** (10k pot incorrectly shown). The correct schematic is presented below.



new products

low-voltage, high-current power supplies



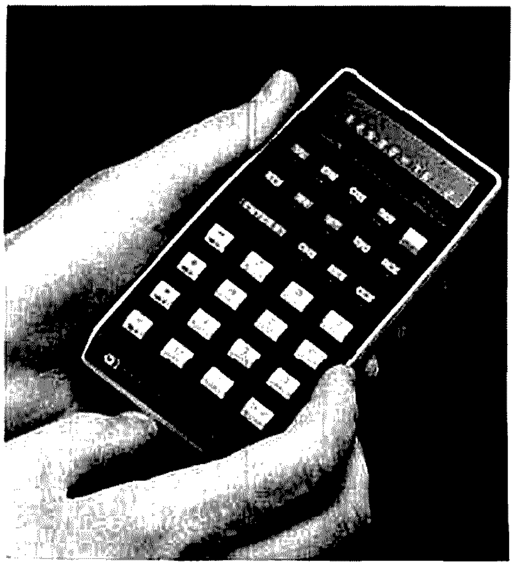
High-current power supplies are an absolute requirement in commercial two-way radio shops and are now on the verge of becoming a requirement in the ham shack. Most of the new solid-state equipment being released for the amateur market is designed for 12-volt operation, either mobile or from a fixed station 12 Vdc supply. The 12-volt ac supply has been a problem in the past because well-regulated, high-current supplies tend to be complex and expensive to build while commercially available supplies are even more expensive and are frequently in short supply. VHF Engineering has recently announced two inexpensive solid-state 12 Vdc power supply kits which are simple to build and can be used in either commercial or amateur applications. Two models are available, the PS12C for 12 amps, and the PS24C for 24 amps.

The circuit for both power supplies consists of a full-wave dc current source feeding a capacitive filter network and an IC voltage regulator. The IC regulator controls a set of pass transistors and keeps the output voltage consistent to within 2% over a load range of from zero to 20 amps (zero to 10 amps in the PS12). Large heat sinks are provided to dissipate the heat produced by the pass transistors. The 12-amp supply is rated at 10 amps continuous or 12 amps for 50% intermittent duty. The 24-amp supply is rated at 20 amps continuous or 24 amps for 50% intermittent duty. Current limiting prevents damage to the supply in the case of an accidental short circuit. The output voltage of both supplies may be adjusted over a nominal range from 12 to 15 volts. The supplies may be used as general purpose, variable voltage supplies by replacing the voltage controlling resistor with a 10k potentiometer.

The VHF Engineering high-current power supply kits are complete with all parts, computer grade capacitor, epoxy glass circuit boards, styled case, and complete instructions. Average construction time is one evening or less. The 12-amp supply kit, model PS12C, is priced at $69.95 ($85.95 wired and tested). The 24-amp supply kit, model PS24C, is $99.95 ($114.95 wired and tested). For more information, write to VHF Engineering, 320 Water Street, Binghamton, New York, 13902, or use *check-off* on page 110.

More Details? CHECK—OFF Page 110

scientific calculator



The new six-ounce HP-21 from Hewlett-Packard is the smallest and lowest priced model in HP's line, and is designed primarily for scientists, engineers and students. The HP-21 has all of the trigonometric and logarithmic functions of the HP-35. In addition, the user can calculate in either degrees or radians; convert from polar to rectangular coordinates and vice versa; format and round the display in either fixed or scientific notation; and perform register arithmetic (+, −, X, ÷) with the contents of the HP-21's single addressable memory.

The new calculator has five fewer keys (30) than other HP pocket models, but since several keys serve dual functions, the HP-21 is able to perform more functions and operations than the HP-35. Like other HP pocket calculators, the HP-21 features the company's RPN logic system with a four-memory stack that holds intermediate answers and automatically brings them back when needed in a calculation.

The HP-21 comes with an owner's handbook, soft carrying case and an ac adapter/recharger that allows the calculator to be operated on ac while its batteries are recharging. Optional accessories include a security cradle and a reserve power pack (with batteries). The HP-21 will be sold through leading col-

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lege bookstores and department stores, by direct mail and through HP's calculator sales force. For more information, use *check-off* on page 110.

525-MHz uhf prescaler

The new Pagel model 525 uhf prescaler divides frequency by ten to extend the range of any 50 MHz or higher counter to the vhf and uhf bands. The unit also contains a 20 dB preamp for the unscaled 1 MHz to 50 MHz range to improve frequency counter sensitivity to 5 millivolts rms or better. Sensitivity is 50 mV rms at 500 MHz, and 30 mV rms below 400 MHz. A through-line feature with an internal signal sampler can be used with transmitters up to 100 watts (requires an external 50-ohm dummy load). This feature can be used to perform simultaneous power and frequency measurements and is a great time saver.

The model 525 operates from the 117 Vac line or battery power (8 to 15 volts) and may be used for portable or mobile use. Price is $159. For more information, write to Pagel Electronics, 6742-C Tampa Avenue, Reseda, California 91335, or use *check-off* on page 110.

envelope detector

The *Modset* recently introduced by David R. Corbin Manufacturing is an envelope detector which can be used with an oscilloscope to provide a clear, hum-free display of the transmitted audio waveform from ssb and a-m transmitters. The *Modset* is built into a rugged, machined aluminum box, and includes input and output impedance matching and an overload protection circuit.

The *Modset* accepts any input signal from a few milliwatts to 25 watts peak power and operates over the frequency

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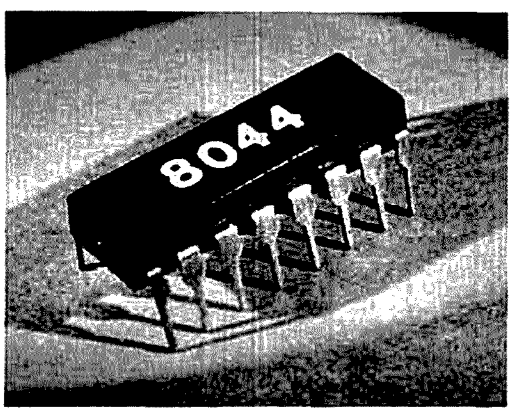
range from 200 kHz to 30 MHz; to 50 MHz with slight reduction in output level. Higher power levels can be monitored by using a short whip antenna or probe instead of the 50-ohm direct coupling. Output level is 0.1 to 10 volts (relative to input power), dc reference plus recovered audio to 10 kHz. The unit is priced at $29.50.

For more information, write to David R. Corbin Manufacturing, Post Office Box 44, North Bend, Oregon 97459, or use *check-off* on page 110.

keyer chip adds dash memory



A companion to the 8043 keyer-on-a-chip has been announced by Curtis Electro Devices. Called the 8044, this new cmos IC offers dash memory in addition to the features found on the 8043. These are dot memory, dot, dash and space completion, instant start, key debouncing filters, iambic operation, internal sidetone, weight control and practically zero power dissipation.

An exact pin-for-pin equivalent to the 8043, the 8044 yields a top performance one-IC electronic keyer capable of running on 5 to 12 volt supplies. Usual power supply is a 9-volt transistor radio battery. The 8044 may be plugged into any keyer designed around the 8043 such as that described in the April, 1975, issue of *ham radio*.

The keyer kits are offered. The 8044-1 contains the 8044, a printed-

circuit card, a socket and instruction manual. The 8044-2 kit contains the foregoing plus the parts necessary to complete a deluxe keyer exclusive of chassis, jacks, switches, speaker and knobs. This keyer will key ±300 volts at 200 mA. The 8044 is priced at $24.95, the 8044-1 kit at $32.95 and the 8044-2 is $57.95. Like the 8043, the 8044 carries a lifetime guarantee.

For further information, contact Curtis Electro Devices, Box 4090, Mountain View, California 94040, or use *check-off* on page 110.

essential formulae for electronic and electrical engineers

This new book by Noel M. Morris provides all of the essential electrical and electronic formulas required by students, technicians and professional engineers. The rapid growth of electronics technology has made it practically impossible to memorize all the formulas which are required. This book contains all the basic equations in the fields of electronics, electrical engineering, control systems, measurements, logic, telecommunications and mathematics. Sections include electrostatics and electromagnetism, complex numbers, ac and dc circuits, transients, amplifiers and oscillators, modulation and transmission lines. The SI system of units is used throughout.

This book assumes that the reader is familiar with each of the formulas, so it does not provide any typical examples or information describing how to use them. However, for the serious worker who frequently needs this information, this valuable book provides it all in one place. 26 pages, softbound, 8¼x11½ inches, $2.95 from Halstead Press, a Division of John Wiley & Sons, 605 Third Avenue, New York, New York 10016.

More Details? CHECK—OFF Page 110

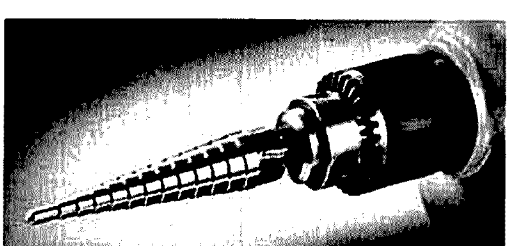
tracking voltage regulators

Three new dual tracking voltage regulator ICs are now available from National Semiconductor. Called the LM125, LM126 and LM127 (LM325, LM326 and LM327 for commercial temperature range), the regulators are designed to provide balanced positive and negative output voltages at currents up to 100 milliamps. Input voltage can be as high as ±30 volts, and there is a provision for external adjustable current limiting.

The LM125 provides tracking outputs of ±15 volts making it ideal for op amp power supplies. It features output voltages balanced to within 1% and line and load regulation of 0.06%. The LM126 provides ±12 volt outputs balanced to within 1% and features line and load regulation of 0.08%, while the LM127 has +5 and −12 volt outputs which are compatible with most mos circuits.

For more information, contact National Semiconductor Corporation, 2900 Semiconductor Drive, Santa Clara, California 95051, or use *check-off* on page 110.

stepped drill bit



The Unibit,® a versatile single flute step drill that does the work of thirteen conventional twist drill bits, is now available. The first Unibit model, designed to fit any three-jawed ¼" (6.5mm) chuck, is intended primarily for use with hand-held electric drills. It has a starting diameter of 1/8" (3mm)

and each of twelve succeeding steps removes material in 1/32" (0.8mm) increments up to 1/2" (12.5mm) diameter. Each step penetrates material up to 1/8" (3mm) thick.

The Unibit is made of industrial grade high-speed steel, heat treated and tempered to assure maximum strength for long wearing, rugged use. It exhibits superior characteristics when drilling thinner gauges of sheet metals such as steel, copper, brass and aluminum as well as most plastics and wood. Starting a hole with the Unibit is a snap because its single flute design eliminates skidding and the need for center punching. Chatter and vibration are also kept to a minimum. Its design geometry also helps prevent the Unibit from penetrating softer material too fast and "hogging-in."

Many other operations, considered difficult with conventional twist drills, can be conveniently accomplished using the Unibit. Its unique cutting edge angle automatically de-burrs a hole, eliminating a time consuming countersink tool change. The Unibit allows for sizing or aligning and reaming operations, drilling intersecting holes as well as enlarging a slot or a square into a round hole. In addition, the Unibit is available with a special non-slip key on its shank that prevents it from spinning in the chuck.

The newest Unibit drill, Model II, enables the drilling of eight round holes, sizes 9/16-inch (14.5mm) to one-inch (25.5mm), with a single bit. The Unibit II requires a starting diameter of 1/2-inch (12.5mm) or larger and removes material in 1/16-inch (1.5mm) increments in eight steps.

Designed for use with any 1/2-inch drill chuck, Unibit II works equally well with hand-held electric drills as well as drill press equipment. The Unibit II is made of industrial grade, high speed steel, heat treated and tempered to assure maximum strength for long wearing use. It's patented design features a single flute to assure smooth penetra-

More Details? CHECK—OFF Page 110

tion of materials and, unlike conventional twist drills, it is easily sharpened without special tools. Unibit II is extremely versatile and allows the user to perform many operations including intersecting holes, making round holes from slots, hole de-burring and a "hole-opening" reamer.

Unibit literature, information and prices are available by contacting the Unibit Corporation, Box 331, Department 2, Wyoming, New York 14591, or by using *check-off* on page 110.

how to use ic logic elements

Just released and completely illustrated, this new book by Jack Streater is designed to help the engineer or technician who has not previously used or designed digital logic circuits meet the challenge of digital ICs in electronics. The practical problems and limitations of connecting IC logic elements into logic systems to accomplish the required result are thoroughly covered.

The first two chapters cover binary, BCD, and decimal number systems and Boolean algebra with its applications to simple switching circuits. The next two chapters discuss gates and gate combinations, and the following chapter explains bistable elements and their uses. Then the logic families (RTL, DTL, TTL, ECL, CTL or CML, MOS and diode logics) are discussed and compared.

Another chapter is devoted to off-the-shelf logic elements — breadboarding, testing, troubleshooting and locating sources of data on integrated circuits. The final chapter includes experiments to aid in understanding the operation of logic circuits. A glossary of digital terms has been included as an appendix.

Soft cover, 160 pages, $4.50 from HR Books, Greenville, New Hampshire 03048.

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ham radio
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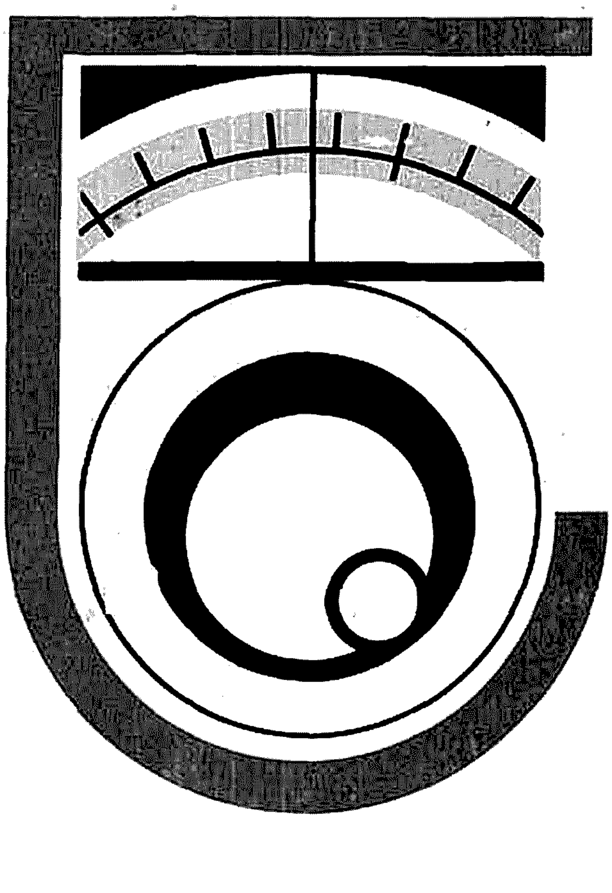
OCTOBER 1975

SPECIAL RECEIVER ISSUE



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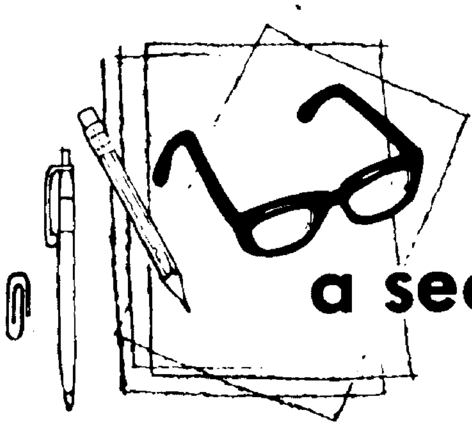

a second look

by jim fisk

As more and more amateurs are discovering, the commercial double-balanced mixer modules which are currently available are the clear choice when you are designing systems which require frequency conversion. The advantages of using these devices (along with sample circuits) were outlined in *ham radio* as long ago as 1968, and more recently in the other amateur radio magazines. Although the double-balanced mixer offers low distortion and high isolation, it suffers from fairly high conversion loss. While single-balanced mixers offer lower conversion loss and can be built less expensively, they do not provide the low distortion and high isolation characteristics of the double-balanced design.

Now, however, engineers at Hewlett-Packard have developed a single-balanced mixer with isolation, distortion and conversion-loss characteristics that equal or exceed those of some double-balanced mixers now on the market. Furthermore, the H-P printed-circuit balanced mixers, part number 5082-9200, are many times smaller (0.57" wide, 0.50" high).

With local-oscillator injection in the range from 200 to 900 MHz, conversion loss of the H-P balanced mixer is about 6.5 to 7.5 dB, a nearly 1 dB improvement over the performance of double-balanced circuits. In addition, isolation is 15 to 20 dB higher, being 45 dB at 200 MHz and 25 dB at 900 MHz. The maximum frequency is specified at 1200 MHz although selected devices would very likely provide good performance on the 1296-MHz amateur band.

In terms of distortion, at a local-oscillator level of +10 dBm (10 mW), the third-order intercept is typically +23 dBm and the 1 dB compression point occurs at +6 dBm. The second-order intercept is at +38 dBm, a roughly 10 dB improvement over conventional double-balanced mixers. Although a filter must be added at some later stage to separate the rf and i-f signals, this is not a particular disadvantage in communications circuits because a selective i-f filter is usually placed in the circuit immediately after the frequency-conversion stage.

Although balanced mixers are never quite as good in practice as they are in theory because of circuit imbalances, the various mismatches often cancel each other to a large degree. To achieve high performance in the new H-P mixer engineers further improved the matching, balancing and symmetry of the circuitry by removing the human element from the manufacturing process. The transformers, for example, are not wound by hand, but are etched on a printed-circuit board with spacing between windings controlled down to 0.001 inch (0.03mm). This technique also balances the parasitic capacitances between windings. With hand winding these elements vary widely as does the overall balance of the circuit.

Since diode matching is important in balancing a mixer, the H-P designers chose their own Schottky process because of its high yield of matched-diode pairs. With the carefully controlled printed-circuit transformers and closely matched diodes, the result is a 1% matching in forward voltages and a 0.1 pF capacitance match.

Although priced at $11 in small quantities, the cost of the H-P mixer drops to less than $2 each in production quantities, so it should see widespread use in commercial applications.

Jim Fisk, W1DTY
editor-in-chief

presstop

THE GETTYSBURG LOG-JAM continues to plague both Amateurs and the FCC as Amateur license applications now are taking up to 12 weeks for final action. This is true even after special efforts (HR Report, August 1) to dig out. Unfortunately, no Amateur tickets were processed while the attempt to catch up on CB applications was underway.

Expansion Plans Provide Hope for the future. Included are a larger, faster computer and a 25% increase in personnel. Helping already are improved procedures and the new box numbers. Remember, address your Amateur applications to Box 1020, Gettysburg, Pennsylvania 17325.

SIX-METER AMATEUR BAND may be in trouble according to Prose Walker's comments at the Texas VHF FM Society's convention. Prose seems to feel that there's a definite push on to add channel 1 to the VHF TV spectrum, and channel 1 is 6 meters!

The Channel 1 agitation may be an offshoot of a proposal made earlier this year by a Dallas consulting firm in response to Docket 20264, which pertains to radio callbox systems in the 72-74 MHz range, suggesting that those four MHz be added to two MHz from the six-meter band to create a new TV channel.

RTTY ENTHUSIASTS are warned that parts for Teletype Models 14, 15, 19 and 20 are going to become harder to get in the near future. Teletype Corp. has announced they'll stop accepting orders for all replacement parts for those models on December 1, 1975, so better order your spares now.

Questions Or Orders can go to Teletype, 5555 Touhy Avenue, Skokie, Illinois 60076, Attn: R.A. Morton, Dept. 3121 — phone (312) 982-2168. Thanks to W8KAJ.

AMSAT QSL BUREAU MANAGER WA1EHF has made a sudden move to California and plans to move the QSL bureau out there as soon as he has a permanent address. In the meantime, WA1EDX will be filling in the void for Dennis and the 288 Grand Street, Bridgeport, Connecticut 06604 address is still valid.

Satellite Use for location of downed aircraft has been proposed, and AMSAT has been asked to participate in feasibility studies. Anyone interested — particularly members in the mid-Atlantic (Washington) area — should contact AMSAT.

OSCAR AND THE HAM, a new video tape copyrighted by the ARRL, has been released for viewing on local PBS stations. Anyone interested should contact their local station and request that they run it.

Help Is Needed in making video tape copies. If you can give any assistance, please get in touch with ARRL Headquarters.

"HOT" HAM GEAR does show up and it pays to be able to prove positive ownership when it does. WB9EBO found his Standard 826 and synthesizer stolen from his home last spring for sale at Hamfesters' hamfest near Chicago, kept an eye on them while W9JZK called the Cook County Sheriff's Police for assistance. Gerry's call, electric pencilled inside the chassis, left the seller (a non-ham who had "bought them from some guy at another hamfest") without an argument.

EUROPEAN BOUND U.S. HAMS should seriously consider taking along a hand-held tuned up for appropriate repeaters on the low end of two meters. W9JUV heard lots of activity in Germany and England on a tunable receiver during a two-week trip in August, and temporary operating permits for many parts of Europe are much easier to get than are their equivalents here.

R. L. DRAKE COMPANY FOUNDER and president Bob Drake passed away July 28th after a long illness. Bob, W8CYE, founded the R.L. Drake Company in 1942 and was active in the successful business until recently. His son Peter has been elected president and promises to continue the traditions of this Amateur-oriented firm. No policy or personnel changes are anticipated in the near future.

JOSEPH JOHNSON, W3GGO, has been named Chief, Rules and Legal Branch of the Amateur and Citizens Division of the FCC effective August 17. Joe replaces John Johnston, K3BNS. This is an important position and we are glad to again see an Amateur in this chair.

receiver noise figure sensitivity and dynamic range — what the numbers mean

A complete discussion of receiver sensitivity, intermodulation distortion, cross modulation and gain compression, and what they mean in terms of performance

James R. Fisk, W1DTY, Ham Radio Magazine, Greenville, New Hampshire 03048

When it came to receivers, the earliest amateur operators were concerned primarily with sensitivity and experimented almost endlessly with different types of crystals, trying to find the one that was the most sensitive. Then came DeForest's Audion, and Armstrong's regenerative detector, and amateurs who could afford the tubes found they had all the sensitivity they could use. However, as the hobby grew, and more and more amateurs started populating the band below 200 kHz, the simple regenerative detector simply wasn't up to the task. Selectivity, with simple tuned input circuits, was practically nonexistent, and the regenerative detector hopelessly overloaded in the presence of strong signals.

In the early 1920s amateurs worked to improve their tuners, but even the so-called "low-loss" tuners were only marginally acceptable. Although several superheterodyne designs were described in the amateur magazines, it wasn't until low-cost, commercial i-f transformers became available in the late twenties that the superhet saw widespread amateur use. Selectivity against interfering signals was still a problem, however, and James Lamb revolutionized receiver design in 1932[1] with his "single-signal" CW circuit which used an i-f stage with extremely high selectivity — provided by regeneration or a simple crystal filter.

The single-signal, single-conversion superhet of the late 1930s suffered from poor rf image response at the higher frequencies, but it wasn't too severe on 14 MHz and few amateur receivers of the day, in fact, tuned much above 18 or 20 MHz (15 meters was not yet assigned to amateur use and most 10-meter operators used specialized receivers or converters). When the 10- and 15-meter bands opened up after the war, however, the poor rf image response of the

single-conversion superhet with a 455-kHz i-f had to be faced – it was solved by going to a double-conversion layout with a first conversion to 2 or 3 MHz to minimize rf image response, and a second conversion to 455 kHz or lower for adjacent channel selectivity.

Although amateur radiotelephone operation in the 1930s was relatively limited, the huge growth of a-m activity after the war demanded improved adjacent-channel phone selectivity. While the crystal filter provided excellent selectivity for CW operation, it was of little or no use on a-m or ssb and some phone operators started using a Q5er – an outboard 80-kHz i-f strip – for improved phone selectivity. This led to the triple-conversion superhets which were the rage of the 1950s.

As pointed out by Goodman,[2] however, the multiple-conversion design had many shortcomings, including the large number of stages between the antenna and the high-selectivity i-f which made it practically impossible to attenuate

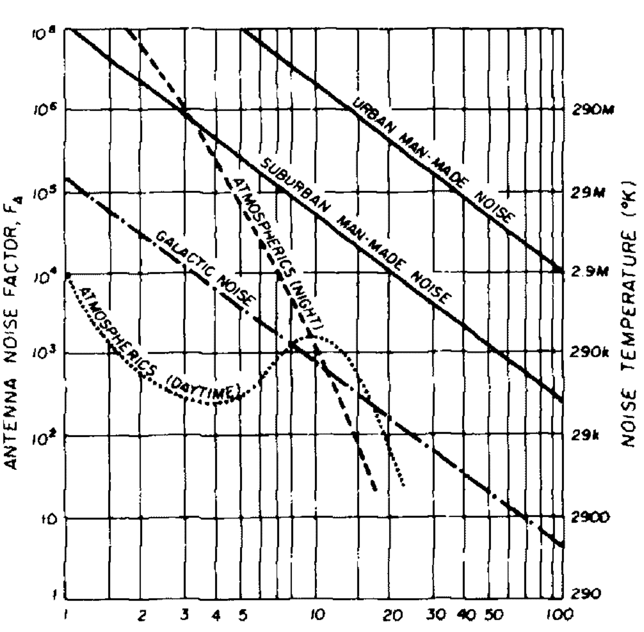

fig. 1. Receiver sensitivity is limited by the external available noise power which varies with frequency. For quiet, rural locations galactic noise is the limiting factor down to about 18 MHz, and atmospheric noise dominates below 18 MHz.

strong, adjacent signals. And, with at least three oscillators running at the same time, it was difficult to avoid the many spurious signals which were generated within the system. He advocated a return to the single-conversion superhet using highly-selective, high-frequency, crystal lattice filters which were just then becoming commercially available.

The 1960s saw a return to single-conversion designs, the use of high-frequency, crystal-lattice filters and the widespread use of semiconductors. With modern devices receiver sensitivity was no longer limited by the rf amplifier (or mixer) stage, but by the external galactic and man-made noise. Cross modulation and overload, on the other hand, were becoming a serious problem as more and more amateurs started using high-power linears and large, directive antennas. Modern communications receivers, therefore, in addition to meeting stringent frequency accuracy, stability, sensitivity, and selectivity requirements, must provide freedom from cross modulation, intermodulation distortion and blocking. Some modern, solid-state solutions to these design goals were discussed recently by Rohde.[3]

The specifications for a typical modern communications receiver might list sensitivity of 0.5 μV for 10 dB signal-plus-noise-to-noise (S+N/N) ratio, intermodulation distortion of −65 dB, "wide" dynamic range, and "virtual elimination" of overload from adjacent signals.

However, is 0.5 μV sensitivity for 10 dB S+N/N adequate for operation on the high-frequency bands? For satellite communications on 10 meters? What is −65 dB intermodulation distortion in terms of signal strength? "Wide dynamic range" and "virtual elimination of overload" are obviously advertising superlatives without definition but what, exactly, can you expect from a high-quality, modern receiver design? Perhaps, if these performance data were defined,

and amateurs understood what they meant, manufacturers would be encouraged to use no-nonsense numerical data. Only then can amateurs compare the dynamic range and cross-modulation performance of one receiver against that of another.

sensitivity

The minimum usable signal or sensitivity of a receiver is determined by the noise in the receiver output. This can be noise generated within the receiver, thermal noise generated by losses in the transmission line, or atmospheric, man-made or galactic noise picked up by the antenna. As shown in **fig. 1**, external noise sources are likely to be the limiting factor up to 100 MHz or so.[4] In urban areas man-made noise predominates and measurements indicate the average level of man-made noise in suburban areas is about 16 dB lower. In a quiet, rural location which has been chosen with care the man-made noise may be near the galactic noise level, but few amateurs are so fortunate.

Atmospheric noise usually predominates in quiet locations at frequencies below about 20 MHz and is produced by lightning discharges so the level depends upon a number of variables including frequency, weather, time of day, season and geographical location. This type of noise is particularly severe during rainy seasons near the equator and generally decreases at the higher latitudes. More complete data on high-frequency atmospheric noise is given in reference 5.

Galactic or cosmic noise is defined as rf noise caused by disturbances which originate outside the earth or its atmosphere. The primary causes of this noise, which extends from 15 MHz well into the microwave region, are the sun and a large number of noise sources distributed chiefly along the Milky Way. Solar noise can vary as much as 40 dB from "quiet" sun levels (low sunspot activity) to periods of "disturbed" sun (high sunspot activity). Galactic noise from the center of the Milky Way is about 10 dB below the noise from a "disturbed" sun, whereas noise levels from other parts of the galaxy can be as much as 20 dB lower. This is important in satellite communications and will be discussed later.

thermal noise

The free electrons in any conductor are in continuous motion — motion that is completely random and is the result of thermal agitation. The effect of this electron motion is to cause minute voltages which vary in a random manner to be developed across the terminals of the conductor. Since this phenomenon was first demonstrated by J. B. Johnson in 1928,[6] thermal noise is sometimes known as Johnson noise. At the same time, H. Nyquist showed, on the basis of the statistical theory of thermodynamics, that the mean square noise voltage generated in any resistance can be expressed as [7]

$$e^2 = 4kTBR \qquad (1)$$

where e^2 = mean square noise voltage
k = Boltzmann's constant = 1.38×10^{-23} joules/°K
T = absolute temperature, °K
B = bandwidth, Hz
R = resistance, ohms

Note that the noise voltage is dependent upon the bandwidth across which it is measured. This implies that noise is evenly distributed across all frequencies which, for all practical purposes, it is.* Although noise bandwidth is not precisely the same as the 3-dB bandwidth of a receiver, in modern receivers with high skirt selectivity the 3-dB bandwidth can be used in **eq. 1** with little error.

The equivalent circuit of any impedance as a source of noise voltage is

*At extremely high frequencies statistical mechanics is no longer valid, and **eq. 1** must be revised on the basis of quantum theory. This equation is valid, however, to at least 6000 GHz.[8]

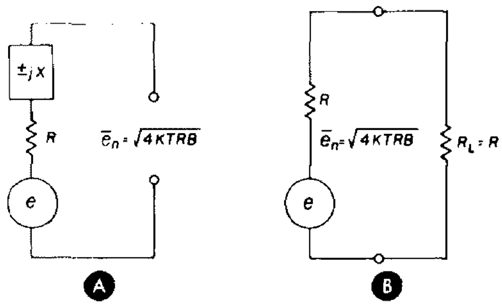

fig. 2. Mean noise voltage depends on temperature, resistance and bandwidth, and is completely independent of reactance as shown in (A). Maximum noise power is transferred to the load when the load resistance is matched to the source resistance (B).

shown in **fig. 2A**. Note that the thermal noise voltage is dependent only on the resistive component and is independent of any reactance in the circuit. As might be expected, maximum noise power is transferred from a thermal noise source when the load impedance presents a conjugate match to the source impedance. This is represented in **fig. 2B** where the load resistance, R_L, is equal to the source resistance. Since $R = R_L$, the noise voltage developed across the load is $e/2$, and from Ohm's law

$$P = \frac{E^2}{R} = \frac{(e/2)^2}{R} = \frac{e^2}{4R} \text{ watts} \quad (2)$$

Substituting the value of e^2 from **eq. 1** into **eq. 2**, the power which can theoretically be transferred under such conditions is called the *available noise power* and is given by

$$P_n = kTB \quad (3)$$

The factor of $4R$ has cancelled out so the available noise power does not depend upon the value of the resistance. This is significant because it means that the available noise power of any resistor (or any noise source), if measured over the same bandwidth, can be represented by a resistor at temperature T. Thus, every noise source has an equivalent noise temperature.

The actual noise power dissipated in the load resistance may be affected by loss in the connecting leads, noise power generated in the load resistor itself, or a less than perfect match to the original resistance. This property is sometimes used in low-noise uhf amplifiers by creating a deliberate (but carefully determined) mismatch between the input termination and the detection device so that something less than the available termination noise power is coupled into the detector.

signal-to-noise ratio and receiver noise figure

The relation of signal amplitude to noise is commonly referred to as the signal-to-noise (S/N) ratio. Unfortunately, this ratio has not been well standardized and is often used interchangeably to mean the ratio of rms signal voltage to rms noise voltage, the ratio of peak signal voltage to peak noise voltage and, in pulse systems, the ratio of peak signal power to average noise power. Therefore, when discussing S/N ratio, it's important to determine exactly which ratio is being referred to.

Although the minimum discernible signal (MDS) that can be heard above the receiver noise level is sometimes used as an indication of receiver sensitivity, it is extremely subjective because it differs many dB from measurement to measurement, and from one operator to another (some experienced weak-signal operators can detect signals as much as 20 dB below the noise level while other operators may have difficulty discerning signals which are equal to the noise level).[9]

Receiver sensitivity has also been defined in terms of a signal-to-noise ratio of unity (signal equals noise)* or equivalent noise floor, but this is difficult to measure unless you have a calibrated signal generator and a spectrum analyzer. Noise figure or noise factor, on the

*This is sometimes erroneously referred to as *tangential* sensitivity. Tangential sensitivity, however, corresponds to a signal-to-noise ratio of 6.25 and is about 8 dB higher.[10]

other hand, is less susceptible to measurement errors than sensitivity and, since its introduction in 1944 by Friis,[11] it has become the accepted figure of merit for receiver sensitivity. Noise figure, NF, is simply noise factor, F, expressed in dB.

$$NF = 10 \log F \, (dB) \qquad (4)$$

The concept of noise factor allows the sensitivity of any amplifier to be compared to an ideal (lossless and noiseless) amplifier which has the same bandwidth and input termination. As far as noise is concerned, that part of a receiver between the antenna and the output of the i-f amplifier can be regarded as an amplifier. The fact that the mixer stage shifts the frequency of the noise does not change the *situation* — it merely causes the noise to lie in a different place in the spectrum from the input noise. The only exception is when the receiver has poor rf image rejection. In this case the noise figure of the receiver is 3 dB worse than it would be if the same receiver had good rf image rejection because the image noise appears at the output along with noise associated with the desired received frequency. This effectively doubles the noise at the output of the i-f amplifier.*

The noise factor, F, of a receiver is defined as

$$F = \frac{S/N \, (ideal\ receiver)}{S/N \, (practical\ receiver)}$$

$$= \frac{S_i/N_i}{S_o/N_o} \qquad (5)$$

S_i = available signal input power
N_i = available noise input power
S_o = available signal output power
N_o = available noise output power

Using this definition, it can be seen that an ideal receiver adds no noise to a signal so its output signal-to-noise ratio is the same as that at the input and the noise factor, $F = 1$.

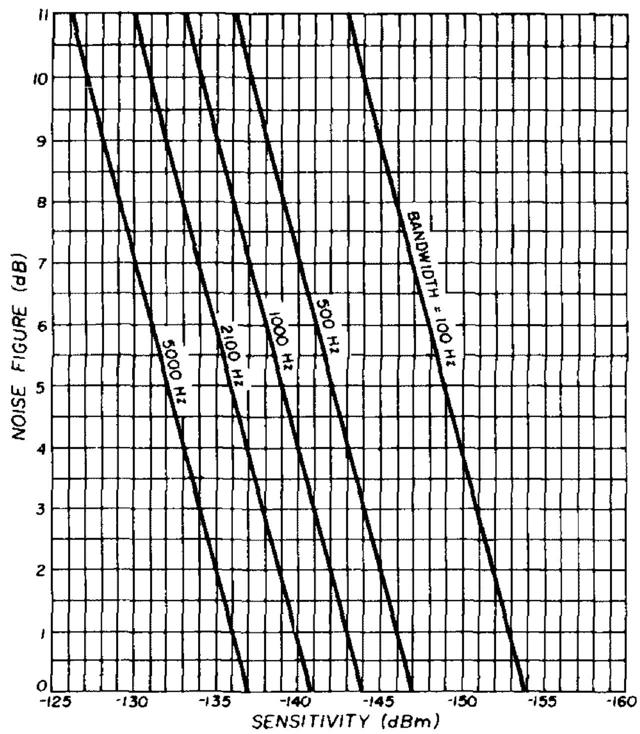

fig. 3. Receiver sensitivity (−dBm) vs receiver noise figure and bandwidth for unity signal-to-noise ratio (signal equals noise). Add 10 dBm for a S/N ratio of 10 dB. Table 1 lists microvolt sensitivity for 50- and 75-ohm systems in terms of dBm sensitivity.

Since the available noise input power, N_i is defined as kT_oB in **eq. 3**, and the power gain of the system, $G = S_o/S_i$, **eq. 5** can be rewritten as

$$F = \frac{N_o}{GkT_oB} \qquad (6)$$

Where T_o is 290° kelvin (IEEE definition). With the receiver noise factor defined in terms of noise output power, N_o, power gain, G, and noise input power, kT_oB, noise factor can be easily correlated to receiver sensitivity. Consider the case where the output signal-to-noise ratio, S_o/N_o, is unity

$$S_o = N_o$$

*The noise figure is always defined at the *input* of the final detector (i-f output) because the noise output of a detector (but not of a mixer) is affected by the presence of a signal. An *fm* signal, for example, will suppress weak noise but will be suppressed itself by strong noise.

For this specialized case, eq. 6 can be rewritten

$$F = \frac{S_i}{kT_oB} \qquad (7)$$

When the temperature is 290°K and the bandwidth is in kHz, $kT_oB = 4 \times 10^{-15}$ mW per kHz. Rewriting eq. 7 in terms of dBm (dB referenced to 1 mW)

$$S_i = 10 \log kT_o + 10 \log B + NF \qquad (8)$$
$$= -144 + 10 \log B_{kHz} + NF (dBm)$$

This function is plotted graphically in fig. 3 for bandwidths commonly found in amateur communications receivers. For example, assume a high-frequency receiver has an 8 dB noise figure at 14.2 MHz and a bandwidth of 2.1 kHz. From eq. 8 or fig. 3, the noise floor of the receiver at 14.2 MHz is at about -133 dBm. An input signal of -123 dBm (10 dB greater) would be required for a 10 dB S/N ratio.* To convert dBm to microvolts, recall that

$$E = \sqrt{RP} \qquad (9)$$

where E is in volts, R is resistance in ohms and P is power in watts. Since -123 dBm is 5.01×10^{-16} watts, -123 dBm is equivalent to 0.16 μV *across* a 50-ohm input termination. However, for a matched signal source, as shown in fig. 4, where the source resistance, R_s, is equal to the load resistance, R_L, the source voltage must be twice the voltage across R_L because of the voltage-dividing effect of the two series resistors in the network. For a *matched* 50-ohm source, therefore, an input signal of -123 dBm requires a source voltage of 0.32 μV. A chart of dBm vs microvolts for *matched* 50- and 75-ohm systems is presented in table 1.

*This is the signal-to-noise ratio in ssb and CW reception. The S/N ratio of a-m and nbfm signals is somewhat less because a-m (and nbfm) detection use only the envelope as a useful output and the S/N ratio must be reduced by a factor which is related to percentage of modulation (or modulation index).

table 1. Microvolt sensitivity vs dBm for *matched* 50- and 75-ohm receiving systems.

dBm	μV 50 ohms	μV 75 ohms	dBm	μV 50 ohms	μV 75 ohms
- 76	70.8	86.7	-111	1.26	1.54
- 77	63.2	77.4	-112	1.12	1.38
- 78	56.2	69.1	-113	1.00	1.23
- 79	50.2	61.5	-114	0.90	1.09
- 80	44.8	54.9	-115	0.80	0.97
- 81	39.8	48.7	-116	0.71	0.87
- 82	35.6	43.6	-117	0.63	0.77
- 83	31.6	38.7	-118	0.56	0.69
- 84	28.2	34.5	-119	0.50	0.62
- 85	25.2	30.9	-120	0.45	0.55
- 86	22.4	27.4	-121	0.40	0.49
- 87	20.0	24.5	-122	0.36	0.44
- 88	17.8	21.8	-123	0.32	0.39
- 89	15.8	19.4	-124	0.28	0.35
- 90	14.2	17.3	-125	0.25	0.31
- 91	12.6	15.4	-126	0.22	0.27
- 92	11.2	13.8	-127	0.20	0.25
- 93	10.0	12.3	-128	0.18	0.22
- 94	9.0	10.9	-129	0.16	0.19
- 95	8.0	9.7	-130	0.14	0.17
- 96	7.1	8.7	-131	0.13	0.15
- 97	6.3	7.7	-132	0.11	0.14
- 98	5.6	6.9	-133	0.10	0.12
- 99	5.0	6.2	-134	0.09	0.11
-100	4.5	5.5	-135	0.08	0.10
-101	4.0	4.9	-136	0.071	0.087
-102	3.6	4.4	-137	0.063	0.077
-103	3.2	3.9	-138	0.056	0.069
-104	2.8	3.5	-139	0.050	0.062
-105	2.5	3.1	-140	0.045	0.055
-106	2.2	2.7	-141	0.040	0.049
-107	2.0	2.5	-142	0.036	0.044
-108	1.8	2.2	-143	0.032	0.039
-109	1.6	1.9	-144	0.028	0.035
-110	1.4	1.7	-145	0.025	0.031

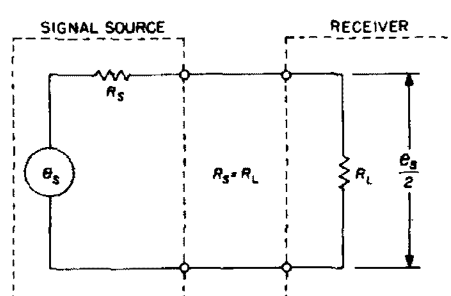

fig. 4. When a signal source is matched to a load, the voltage across the load is one-half the source voltage because of the voltage-dividing effect of the source and load resistors. When making receiver sensitivity measurements in the laboratory, a 6-dB attenuator is placed in the line so sensitivity can be read directly from the signal-generator's internal calibrated attenuator.

Because of this two-to-one voltage-dividing effect, you must be very careful when comparing the sensitivity of one receiver against that of another. An *input* of +119 dBm, for example, implies an input directly at the receiver terminals and is 0.25 μV rms across 50 ohms. Sensitivity, on the other hand, implies the use of a matched signal generator so *sensitivity* of 0.25 μV corresponds to an *input* of −125 dBm. This is a 6 dB difference. Since most amateur receiver manufacturers tend to use *sensitivity* specifications, there is no advantage, but the difference must be considered when you calculate the receiver noise figure and dynamic range. In this article inputs will be stated in dBm as this eliminates the 6 dB *conversion factor* — **table 1** can be used to convert to *sensitivity*.*

cascaded stages

A relatively simple equation for the noise factor of a receiving system, in terms of the individual stage gains and noise factors is

$$F_T = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} \qquad \textbf{(10)}$$

F_T = overall system noise factor
F_1 = noise factor of the first stage
F_2 = noise factor of the second stage
F_3 = noise factor of the third stage
G_1 = power gain of the first stage
G_2 = power gain of the second stage
G_3 = power gain of the third stage

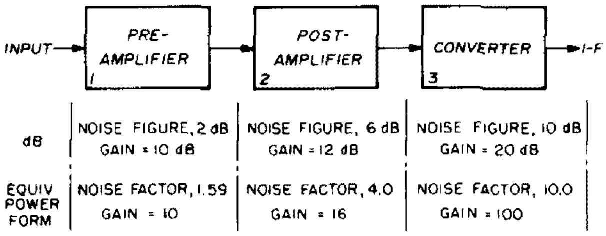

fig. 5. Noise factor of cascaded stages can be calculated by using equation 10 where all quantities are in power ratios. Noise factor at the input of the three cascaded stages shown here is 1.946 (noise figure = 2.9 dB).

Note that all terms are in power ratios. If the gain of the first stage is high, and the noise factor of the second stage is low, the overall system noise factor is determined primarily by the first stage and the third term of **eq. 10** may be dropped.

For example, consider the block diagram of the first three stages of a typical vhf receiving system shown in **fig. 5.** The overall system noise factor is

$$F_T = 1.59 + \frac{4-1}{10} + \frac{10-1}{10 \cdot 16}$$

$$= 1.59 + 0.3 + 0.056 = 1.946$$

$$NF_T = 10 \log 1.946 \approx 2.9\ dB$$

For this receiver the overall system *noise figure is* 0.9 dB *higher than that of* the preamplifier alone. Raising the preamplifier gain to 13 dB or dropping the noise figure of the second stage to 4 dB (F = 2.5) would reduce the system noise figure to approximately 2.45 dB (F = 1.75). Depending on the frequency and the application, this may be a worthwhile improvement.

noise temperature

It is often convenient when working with very low noise uhf and microwave receivers to represent the noise figure of the receiver as an equivalent noise temperature. This is because the noise temperature of a receiving system varies over the range from 0° to 174°K as the noise factor varies from 1.0 to 1.6 (0 to 2 dB noise figure) and noise calculations using equivalent noise temperatures will provide better accuracy.

As mentioned above, the noise figure of an ideal receiver is 1, so the component of receiver noise figure which is due to internally generated noise, N_r, is

*The usual procedure for measuring receiver sensitivity is to place a 6 dB attenuator between the signal generator and the receiver. Receiver sensitivity can then be read directly from the signal generator's calibrated output attenuator.

$F - 1$. Therefore

$$N_r = (F - 1) G k T_o B \qquad (11)$$

The internally generated noise, N_r, can be represented by a noise power, GkT_rB, where T_r is the equivalent noise temperature. Substituting into **eq. 11**, this equivalent noise temperature can be expressed in terms of the reference noise temperature, T_o = 290°K.

$$T_r = (F - 1) T_o \qquad (12)$$

Eq. 12 can be easily rearranged to express noise factor as a function of the receiver noise temperature, T_r, and the reference noise temperature, T_o

$$F = 1 + \frac{T_r}{T_o} \qquad (13)$$

Although noise temperature is seldom used by amateurs, it is a more basic unit than noise factor and is actually easier to deal with, both in understanding concepts and making practical noise calculations. For a more complete discussion of noise temperature, see reference **12**.

transmission lines

The transmission line's contributions to receiver noise come from a common source: line losses. The first of these is the more obvious. When a signal travels down a lossy transmission line, the signal is attenuated. This reduces the signal-to-noise ratio and is equivalent to increasing the noise factor of the receiver. This increase can be calculated by introducing a loss factor, L, which is the loss of the cable expressed as a power ratio.

The second effect is due to the noise factor of the transmission line. The fact that the line has losses implies that there is a loss resistance associated with it (which is distinct from characteristic impedance). Since the line is warm it generates noise due to thermal agitation. The noise factor of the line, F_t, is related to the loss factor, L, and the physical temperature of the line, T_t, by the following equation

$$F_t = \frac{(\frac{1}{L} - 1) T_t}{290} + 1 \qquad (14)$$

The degradation of the receiver noise figure due to transmission line contributions may be calculated by considering the transmission line and the receiver as cascaded stages and using a form of **eq. 10**

$$F_{tr} = F_t + \frac{F_r - 1}{L} \qquad (15)$$

where F_{tr} = noise factor of the receiver and transmission line
F_r = noise factor of the receiver
F_t = noise factor of the transmission line
L = loss factor of the transmission line

For example, a receiver with a noise factor of 4 (NF = 6 dB) is used with a transmission line which has a loss factor of 0.63 (2 dB loss). The physical temperature of the line is 300°K (F_t = 1.61). What is the combined noise figure?

$$F_{tr} = 1.61 + \frac{4 - 1}{0.63} = 6.37$$

$$NF_{tr} = 8.04 \ dB$$

When line losses are low but receiver noise figures are 3 dB or greater, line loss is the predominate contributor to increased noise figure. When receiver noise figure is very low, the thermal effect predominates.

antenna noise

Of all the contributions to system noise, antenna noise is probably the least understood. Assuming the antenna is built of good conducting materials, it contributes virtually no thermal noise of its own to the receiving system. The noise power the antenna does deliver to the receiver depends almost entirely on

table 2. Performance of a receiver with 0.5 μV sensitivity for 10 dB S+N/N with 100 feet (30.5m) of RG-8A/U transmission line is shown in first two columns. Third column lists external available noise power for quiet receiving locations on each of the amateur bands. Fourth column shows receiver signal (50-ohms) required for 10 dB S+N/N on each of the bands (based on external noise). Last column lists acceptable noise figure for each of the bands (see text). Bandwidth = 2.1 kHz.

frequency	noise factor at antenna	noise figure	external available noise power	receiver input signal for 10 dB S+N/N	acceptable noise figure
1.8 MHz	15.8	12.0	- 93 dBm	15.3 μV	45 dB
3.5 MHz	16.2	12.1	-101 dBm	12.6 μV	37 dB
7.0 MHz	16.7	12.2	-111 dBm	4.0 μV	27 dB
14.0 MHz	17.6	12.5	-113 dBm	3.1 μV	24 dB
21.0 MHz	18.3	12.6	-118 dBm	1.8 μV	20 dB
28.0 MHz	18.9	12.8	-123 dBm	1.0 μV	15 dB
50.0 MHz	20.9	13.2	-129 dBm	0.5 μV	9 dB
144.0 MHz	26.9	14.2	-139 dBm	0.2 μV	2 dB

the temperature and other physical characteristics of the material lying in the antenna's field of view. A 432-MHz moonbounce receiving antenna looking out into space, for example, may deliver only as much noise power as a resistor at 10°K (noise factor = 1.03). If this same antenna is rotated so that the warm earth comes into its field of view, the antenna noise temperature would rise to about 300°K (noise factor = 2).

A ten-meter Oscar receiving antenna which is pointed into "cold" space, on the other hand, will see a noise temperature of about 10,000°K minimum (noise factor = 35). Pointed at the horizon, however, the antenna noise temperature may be ten or fifteen times higher, depending upon the amount of man-made noise.

There is little that can be done to improve the situation of a terrestial radio circuit, but an antenna that looks at the sky, such as a satellite antenna, deserves careful design. This is because the effect of the earth is still present, and any sidelobe that sees the earth will pick up thermal noise. This is sometimes quite serious and sidelobes are of major concern in many deep-space communications and radio astronomy systems. Careful attention to antenna design with respect to sidelobes can provide antenna temperatures significantly under 50°K, while poor design can result in much higher values.

minimum usable sensitivity

With an understanding of receiver noise factor and its relationship to signal-to-noise ratio, it's now possible to determine the *minimum usable sensitivity* (MUS) of a receiving system, and how the performance of your own equipment affects your ability to receive weak signals. Let's first consider that modern communications receiver mentioned earlier which had a specified sensitivity of 0.5 μV for 10 dB S+N/N ratio.*

From **table 1** a 0.5 μV *sensitivity* for 10 dB S+N/N is equivalent to a sensitivity of −119 dBm in a 50-ohm system (−129 dBm noise floor). Assuming a bandwidth of 2.1 kHz, the noise figure of the receiver is about 11.8 dB (noise factor = 15.1). Assuming 100 feet

*A S+N/N ratio of 10 dB corresponds to a S/N ratio of 9.54 dB, a difference small enough to be neglected for all practical purposes. A S+N/N ratio of 3.5 dB, however, corresponds to a S/N ratio of about 0.9 dB, and the difference must be considered. S+N/N may be converted to S/N by using the relationship [(S+N/N) − 1] = S/N (in power ratios, *not* dB).

(30.5m) of RG-8A/U transmission line at 300°K, the noise factor at the antenna terminals may be calculated from eq. 15, and is shown in table 2 for the six high-frequency amateur bands, (calculated on the basis of 0.5 μV sensitivity for 10 dB S+N/N on all bands, which may be optimistic). Even at 28 MHz, where the line loss has increased the noise factor by 25 per cent, the system noise factor is still well below the available noise power seen by the antenna (see fig. 1).

At 50 MHz, however, the system is limited by receiver noise and a lower noise figure would be required for weak-signal work (system noise at 50 MHz in this example is about 2.4 dB higher greater than external noise for a quiet location). If the receiver was connected directly to the antenna terminals to eliminate transmission line losses the system noise figure would be essentially that of the receiver alone and a 0.5 μV signal would provide the desired 10 dB S+N/N ratio. Although 100 feet (30.5m) of RG-8A/U coaxial cable has only about 1.35 dB loss at 50 MHz, it degrades the noise figure sufficiently that the system is no longer limited by external noise sources. This points up the importance of using low-loss transmission lines (or mounting a receiving preamp at the antenna).

Assuming a quiet, rural location that is limited primarily by galactic noise down to about 18 MHz, and atmospheric noise below 18 MHz, what is the *minimum* usable receiver sensitivity for terrestial communications?

As can be seen from table 2, rather poor receiver sensitivity is acceptable on 40, 80 and 160 meters because the external noise at these frequencies is very high. This also explains why the simple receivers of the 1920s were relatively successful. The high external noise levels also make it possible to use rather inefficient receiving antennas on the lower frequencies.[13] It's important to note that a 0.5 μV signal is of little practical use on 160, 80 or 40 meters because it would be buried in the noise level.

The required sensitivity on 20, 15 and 10 is not difficult to obtain with modern devices, but receivers which are optimized for the lower frequencies may not offer top performance on 10 meters. It should be pointed out that the "acceptable" noise figure in the last column of table 2, is somewhat arbitrary and is based on setting the receiver noise floor about 3 dB below the external noise floor. This is probably adequate 90 per cent of the time, but since noise varies randomly, a statistical analysis indicates there may be times when a lower noise figure may be desirable. However, it is generally agreed that a 10 dB noise figure is more than adequate up to 22 MHz and an 8-dB noise figure may occasionally prove useful on 10 meters. Why design a high-frequency receiver for extraordinary sensitivity when its performance is limited by external noise over which you have no control? A very sensitive receiver is more prone to intermodulation and cross-modulation effects, and these may be more important.

At vhf the external noise levels are much lower and low-noise receivers are required for good weak-signal performance. Since it's relatively easy to build low-noise receivers for 50 MHz with modern semiconductors, there's no excuse for being limited by system noise figure on this band. A receiver with a 5 dB noise figure at 50 MHz, for example, when used with 100 feet (30.5m) of RG-8A/U transmission line, will provide a system noise factor of 4.38 (NF = 6.4 dB) at the antenna terminals. This is well below the external noise.

The 144-MHz example in table 2 is hopelessly inadequate and represents at least 12 dB degradation over what can be obtained in practice. A receiver with a 1.5 dB noise figure on this band, when used with the 100 feet (30.5m) of RG-8A/U, will provide a system noise

factor of 2.54 (NF = 4.1 dB) at the antenna terminals which is still inadequate. A transmission line with 0.7 dB loss would bring receiver noise figure within acceptable limits, but it might be easier and less expensive to install a low-noise preamp at the antenna.

As pointed out earlier, the noise temperatures of antennas that are pointed into space for satellite communications (or EME) are much lower than for terrestial communications where the antenna is pointed at the horizon. This means that the receiver noise figures must be lower for maximum performance. Some parts of the sky are noisier than others due to the presence of noise sources, but the noise figure of the receiver should ideally be low enough that the system is galactic-noise limited. Following are the receiver noise figures to shoot for when designing receivers or converters for satellite communications on vhf.

frequency	galactic noise floor	noise figure
28 MHz	−125 dBm	8 dB
50 MHz	−130 dBm	5 dB
144 MHz	−139 dBm	1 dB
220 MHz	−140 dBm	0.7 dB
432 MHz	−141 dBm	0.2 dB

These figures are based on a 2.1-kHz bandwidth and assume a lossless transmission line. For more accurate calculations at low noise figures, the use of noise temperatures is recommended.

Although the topic of noise figure measurement is beyond the scope of this article, the simplest method of making the measurement is to compare receiver noise to the noise generated by a temperature-limited vacuum diode. This technique is easily applied in the home workshop and has been discussed many times in the amateur radio magazines.[14,15,16] Guentzler also described a noise measuring system which used a pilot lamp as a noise source.[17]

intermodulation distortion

Amplitude distortion occurs in an amplifier when the magnitude of the output signal is not exactly proportional to the input signal. Although amplifiers can be designed to be nearly perfectly linear over a portion of their operating range, every amplifier has nonlinearity which can cause distortion products or harmonics of the driving waveform. *Intermodulation distortion* or IMD is a type of amplitude distortion which occurs when a nonlinear amplifier is driven by more than one discrete frequency. Although the discussion here is limited to IMD in receivers, this is the same distortion which is used to define the linearity of ssb linear power amplifiers.

When an rf signal with varying amplitude is passed through a nonlinear device, many new products are generated. The frequency and amplitude of each component can be calculated mathematically since the nonlinear device can be represented by a power series expanded about the zero-signal operating point.[18] Although many products are generated, the ones of primary concern are the second and third. This can be demonstrated with a two-tone signal with outputs at f_1 and f_2, 14001 and 14003 kHz.

$$\begin{aligned} f_1 &= 14001 & f_2 &= 14002 \\ 2f_1 &= 28002 & 2f_2 &= 28004 \\ 3f_1 &= 42003 & 3f_2 &= 42006 \end{aligned}$$

Although each of the harmonics fall well outside the passband of a receiver which is tuned to pass 14001 and 14002 kHz, the harmonics mix together to produce intermodulation products which do fall within the passband. The third-order products consist of

$$\begin{aligned} 2f_1 - f_2 &= 14000 \text{ kHz} \\ 2f_2 - f_1 &= 14003 \text{ kHz} \end{aligned}$$

The fifth-order products consist of

$$\begin{aligned} 3f_1 - 2f_2 &= 13999 \text{ kHz} \\ 3f_2 - 2f_1 &= 14004 \text{ kHz} \end{aligned}$$

The output spectrum is shown in **fig. 6**. Unless the nonlinearity of the amplifier

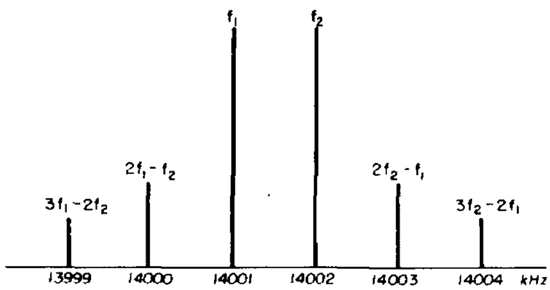

fig. 6. Third- and fifth-order intermodulation products generated by input signals at 14001 and 14002 kHz. In receiver stages fifth-order IMD is usually small enough to be neglected.

is particularly severe, fifth-order IMD is not usually a problem and can be ignored in receiver applications.

Although the IMD distortion products which are generated by two discrete frequencies are used here because they're easy to visualize, exactly the same sort of thing occurs with complex speech waveforms.

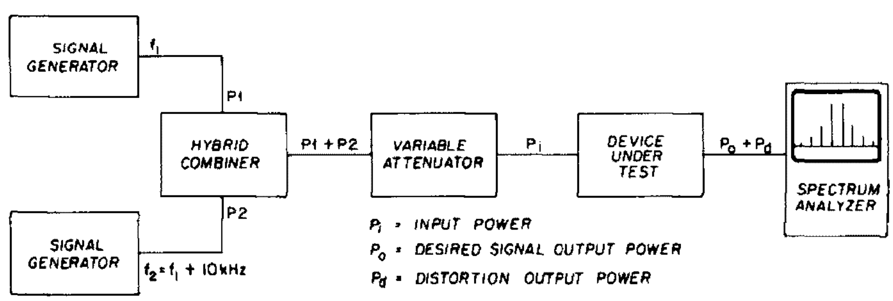

fig. 7. Block diagram of the test setup for evaluating the IMD performance of an amplifier or mixer with a spectrum analyzer.

In a receiver rf amplifier or mixer stage IMD may be caused by two adjacent CW signals or by a ssb signal. Furthermore, in a mixer where the input must be wideband (such as the double-balanced mixer which is currently finding wide use), two input signals, (f_{R1} and f_{R2}) may mix with the local oscillator (f_L) to produce in-band, two-tone, third-order intermodulation products $(2f_{R1} - f_{R2}) \pm f_L$ and $(2f_{R2} - f_{R1}) \pm f_L$. Third-order intermodulation products also occur at $(2f_{R1} - f_{R2})$ and $(2f_{R2} - f_{R1})$. Two input frequencies at 14210 and 14230 kHz, for example, with a 5.2-MHz local oscillator (9 MHz i-f), will produce two-tone, third-order intermodulation products at 8990 and 9050 kHz (i-f passband) and 14190 and 14250 kHz (rf passband).

Third-order IMD is measured in the laboratory with a spectrum analyzer using the test setup in fig. 7. However, the concept of the third-order intercept is finding increased use to describe the IMD response of mixers, and can also be used to describe the linearity of amplifiers.* The third-order intercept is the theoretical point where the two-tone, third-order response is exactly equal to the two-tone input. Amplifiers and mixers are not operated at this level in practice, but the intercept point offers an internationally recognized figure of merit for comparison of devices, both active and passive.

In addition, the intercept point permits comparison of amplifiers and mixers where the intermodulation specifications are given at different two-tone levels. Once the intercept point is known, you can calculate the two-tone, third-order response at any input level by simply remembering that every 1-dB change in the two-tone input produces a 3-dB change in the third-order output. With this information it is possible to predict the maximum rf input level which is allowable.

With each 1 dB decrease in the f_R input level, for example, the third-order

*Class A amplifiers. Linear class AB or B amplifiers often exhibit two-tone, third-order intermodulation products which follow an S-shaped curve that both increases *and* decreases with additional input signal level so they cannot be compared by the intercept point method.

product is decreased an *additional* 2 dB. As shown in fig. 8, a high-level double-balanced mixer will suppress third-order products about 65 dB when both signals are at zero dBm (224 mV across 50 ohms) and 85 dB when both input signals are at −10 dBm (71 mV across 50 ohms). The third-order intercept point for these mixers is +27.5 dBm, relative to the *output*. Relative to the *input*, the intercept point is at +32.5 dBm. This is 17 dB higher than the intercept point for a low-level double-balanced mixer such as the Minilabs SR1A or Anzac MD108. The 3-dB compression point shown on the graph is a combination of both conversion compression and desensitization.

intercept point

The third-order intercept point, IP, can be calculated from the relationship

$$IP = \tfrac{1}{2}(P_o - P_d) + P_i \qquad \textbf{(16)}$$

IP = third-order intercept, dBm
P_o = desired output, dBm
P_d = third-order distortion products, dBm
P_i = input power, dBm

Since the third-order IMD is defined as $(P_o - P_d)$, **eq. 16** can be rewritten as

$$IP = \tfrac{1}{2}IMD + P_i \qquad \textbf{(17)}$$

For the spectrum display shown in **fig. 9**, for example, the third-order intermodulation distortion products with two input signals of 4 mV (−35 dBm) are 80 dB down, and the third-order intercept is

$$IP = 0.5 \cdot 80 - 35 = +5\ dBm$$

Most amateurs don't have spectrum analyzers, but if good intermodulation distortion *information* is provided on the receiver data sheet, the intercept point can be easily calculated with **eq. 17** (in all too many cases, however, amateur receiver manufacturers ignore IMD completely, and when they do provide IMD data, it is incomplete). How-

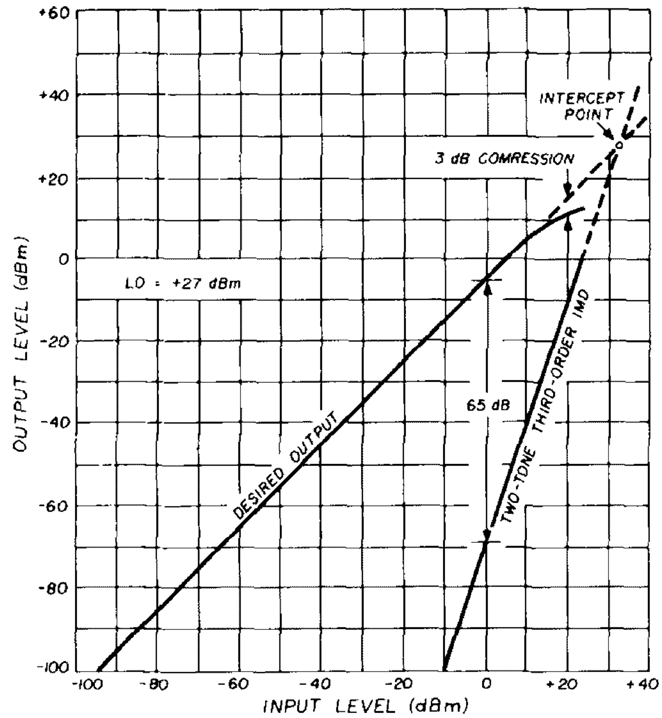

fig. 8. Third-order intercept point of high-level double-balanced mixer (SRA1H) at 50 MHz is +32.5 dBm relative to the input, 3-dB compression occurs at +20 dBm and third-order IMD is suppressed 65 dB when both input signals are 0 dBm (224 mV across 50 ohms).

ever, assume the specifications for an amateur-band receiver list −75 dB IMD for an input of 1 mV (−47 dBm). The third-order intercept is

$$IP = 0.5 \cdot 75 - 47 = -9.5\ dBm$$

Once the intercept point is known, the IMD performance at any input level can be found by rearranging **eq. 17**.

$$IMD = 2(IP - P_i) \qquad \textbf{(18)}$$

For an intercept point of −9.5 dBm, for example, the IMD at various input levels is shown below.

input signal	IMD
100 μV (−67 dBm)	115 dB
500 μV (−53 dBm)	87 dB
1000 μV (−47 dBm)	75 dB
5 mV (−33 dBm)	47 dB
10 mV (−27 dBm)	35 dB

Compare this with the state-of-the-art receiver front end described on page 26 of this issue which has −74 dB IMD at an input of 100 mV (−7 dBm). The

intercept point is at +30 dBm.

input signal	IMD
100 μV (−67 dBm)	194 dB
500 μV (−53 dBm)	166 dB
1000 μV (−47 dBm)	154 dB
5 mV (−33 dBm)	126 dB
10 mV (−27 dBm)	114 dB

The superiority of this receiver is obvious — for all but the strongest signals the IMD products are at or below the receiver noise level. Assuming a 10 dB noise figure and 2.1-kHz bandwidth, an input signal of −23.7 dBm (14.7 mV across 50 ohms) will produce IMD products just equal to the noise level. In the receiver with an intercept point at −9.5 dBm, however, the IMD products are already 3 dB greater than the noise with an input of −49 dBm (800 μV). This will be discussed further under the subject of dynamic range.

double-balanced mixers

Often a mixer data sheet does not specify the third-order intercept point, but a rule-of-thumb estimate can be easily made by examining the 1-dB compression point. As the rf input is increased, the i-f output should follow in a linear manner. However, after a certain point, the i-f output increases at a lower rate until the mixer output becomes fairly constant. The point at which the i-f output deviates from the linear curve by 1 dB is called the 1-dB compression point. At this point the conversion loss is 1 dB greater than it was when the rf input was smaller.

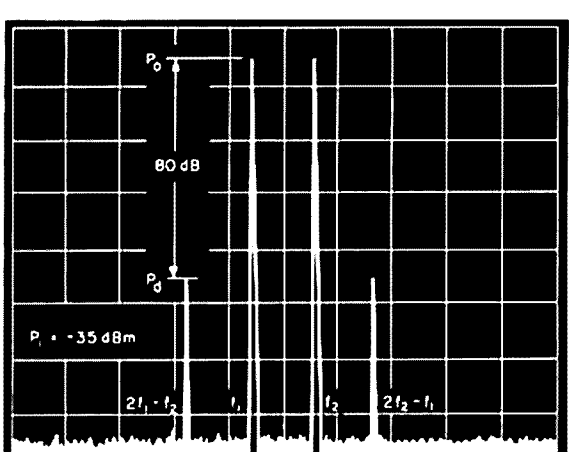

fig. 9. Spectrum analyzer display of communications receiver shows IMD is 80 dB down with a two-tone input of −35 dBm (4 mV across 50 ohms). Intercept point is at +5 dBm.

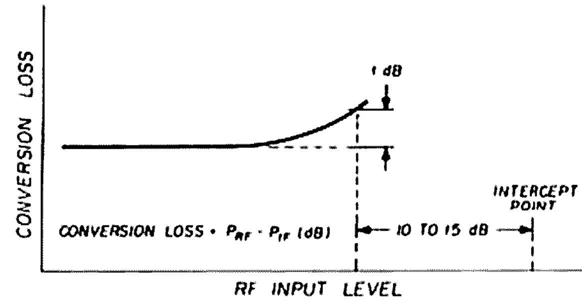

fig. 10. When working with double-balanced mixers, the third-order intercept point may be estimated by using the rule of thumb that the intercept point is 10 to 15 dB above the 1-dB compression point (see text).

The importance of the 1-dB compression point is its utility in comparing the dynamic range, maximum output and two-tone performance of various double-balanced mixers. As a rule of thumb, the third-order intercept point is approximately 10 to 15 dB higher than the 1-dB compression point[19] (about 15 dB at the low frequencies and 10 dB at higher frequencies). This is shown in fig. 10.

To properly use a double-balanced mixer, it is necessary to relate the two-tone input and third-order output levels to avoid generating excessive distortion which would compromise the final design. This is equally valid for amplifiers. Also important, but not as obvious, is the effect higher operating frequencies have on the double-balanced mixer's two-tone, third-order distortion characteristics. Performance is usually better at the lower frequencies and drops off as frequency is increased. For typical high-frequency double-balanced mixers with a maximum frequency specification of 500 MHz, performance starts to fall off somewhere between 50 and 100 MHz.

As discussed by DJ2LR on page 26, it is now possible, using a double-

balanced mixer in the front end, to build high-frequency communications receivers with a third-order intercept at +30 dBm. Using the rule of thumb that the 1-dB compression point is 15 dB below the intercept point, 1-dB compression occurs at an input of approximately +15 dBm or 1.25 volts across 50 ohms.

By comparison, the 1 dB compression point of many commercial amateur receivers is in the vicinity of −20 dBm (22 mV across 50 ohms) and some solid-state receivers with bipolar rf amplifiers go into compression at −40 dBm (2 mV across 50 ohms).

cross modulation

Another type of amplitude distortion which can occur in tuned amplifiers is *cross modulation*. This is related to IMD and is produced when the modulation from an undesired signal is partially transferred to a desired signal in the passband of the receiver. The 3-dB compression point in **fig. 8** describes the start of cross-modulation effects.

The cross-modulation effect is independent of the desired signal level and is proportional to the square of the undesired signal amplitude. Because of this relationship, an rf attenuator which lowers the signal level at the input to the receiver may provide a great improvement in cross-modulation performance. A 6 dB attenuator at the receiver input terminals, for example, will reduce cross modulation by 12 dB. If the desired signal is at least 6 dB above the level at which the receiver provides a satisfactory S/N ratio this results in a marked improvement in received signal quality.

Cross modulation is measured in the laboratory by setting one signal generator to deliver a CW output and another generator is set up for 30% amplitude modulation. The output of the a-m generator is increased until 1% modulation appears on the CW signal as measured with a spectrum analyzer. This represents a cross-modulation ratio of about 30 dB (cross-modulation level 30 dB below the reference level).

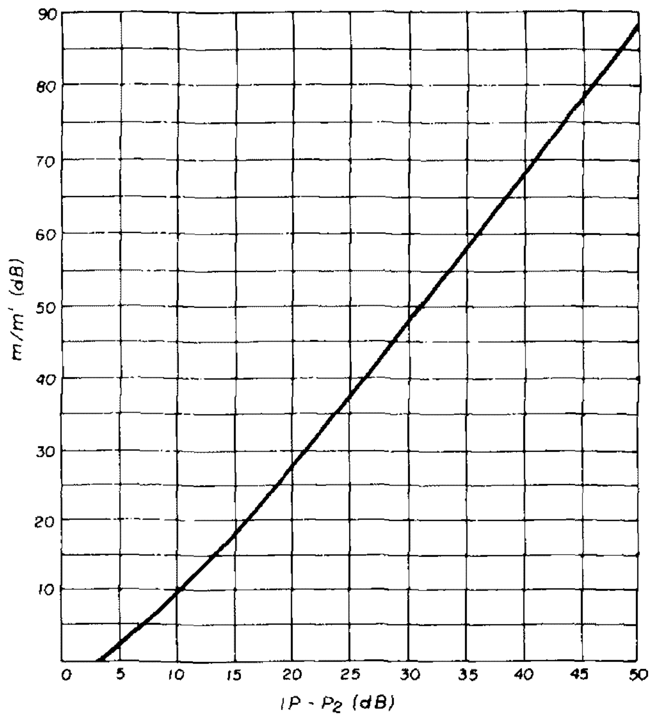

fig. 11. Graph of cross modulation vs the ratio of the intercept point to the interfering signal level in dB.

Cross modulation is related to the intercept point by the relationship

$$m/m' = (P_{ip}/4P_c) - ½ \quad \textbf{(19)}$$

m/m' = ratio of cross modulation transferred from a large signal to a smaller one

P_{ip} = intercept point power

P_c = interfering signal power

Cross modulation in dB is simply

$$m/m'(dB) = 20 \log(m/m') \quad \textbf{(20)}$$

In **fig. 11** cross modulation is plotted against the difference between the intercept point and the cross-modulating signal in dBm. Cross modulation of 30 dB, for example, corresponds to 21 dB difference between the intercept point and the signal producing the cross modulation. For a receiver with an intercept point at +30 dBm, a modulated input signal at a level of +9 dBm (630 mV

across 50 ohms) will produce 30 dB cross modulation. For a receiver with an intercept point at −9.5 dBm (the more usual case), an interfering signal level of −30.5 dBm (6.7 mV) will produce 30 dB cross modulation.

gain compression

When a receiver is tuned to a weak signal, a strong, adjacent signal may cause an apparent decrease in receiver gain. This is called *compression* or *desensitization* and occurs when the input voltage from the undesired signal is large enough to exceed the bias on an rf amplifier or mixer and drives the base (or grid) into conduction. This reduces gain, as shown by the compressed curve of **fig. 8**, and increases distortion. The rectified base (or grid) current can also be coupled back to the receiver's agc system which results in a further reduction in overall receiver gain.

Compression is measured by setting one signal generator to produce a CW signal and another generator, at a given frequency spacing, is adjusted to depress the desired signal a certain amount, usually 3 dB. Like cross modulation, however, a compression specification has little meaning if the frequency separation between the two signals is not specified.

Since both cross modulation and compression are caused by strong, undesired signals which are adjacent to the receiver passband, they can be controlled to a certain extent by the selectivity at the front end of the receiver. High receiver sensitivity, of course, is the antithesis of good cross-modulation and compression performance — this reinforces the argument for receiver noise figures on the order of 10 dB for the high-frequency range.

dynamic range

The front end of a receiver is subjected to a multiplicity of input signals which tend to intermodulate to produce a level of distortion products which is dependent on the magnitude of the incoming signals. Therefore, the upper end of a receiver's dynamic range is defined by the input signal level which produces third-order IMD products just equal to the receiver's noise level.

At the lower end dynamic range is limited ultimately by the noise figure of the receiver. Also important, however, is the way the receiver handles weak signals. Some linear rf amplifiers and mixers give good performance in the middle of their operating range but exhibit transfer curves that introduce considerable distortion at low signal levels. With proper design, however, this is not a problem, and the dynamic range of a receiver is usually defined as the *spurious-free dynamic range* where the maximum input signal is as defined above and the minimum input signal is at the noise floor of the receiver (**eq. 8**).

$$DR = 2/3\,(IP - N_o) \qquad \textbf{(21)}$$

where DR = spurious-free dynamic range, dB
IP = intercept point, dBm
N_o = receiver noise floor, dBm

The dynamic range of a receiver is important because it allows you to directly compare the strong-signal performance of one receiver against that of another. On today's crowded bands, and the high incidence of kilowatt transmitters and directive antennas, strong-signal performance is usually much more important than sensitivity. Although dynamic range can be used as a figure of merit, it's also useful to know the maximum input signal level, $P_{i(max)}$, which will produce third-order IMD products just equal to the noise level. This can be calculated from*

$$P_{i\,(max)} = 1/3(2IP + N_o) \qquad \textbf{(22)}$$

$P_{i\,(max)}$ = maximum input signal, dBm
IP = intercept point, dBm
N_o = receiver noise floor, dBm

* Derivation of **eq. 16** through **eq. 22** will be sent to interested readers upon receipt of a self-addressed, stamped envelope.

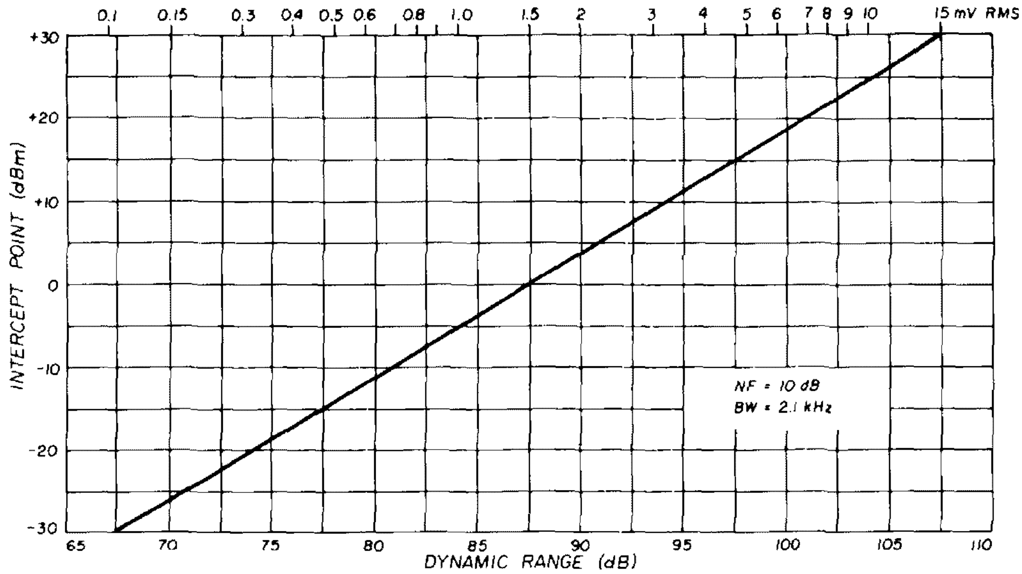

fig. 12. Spurious-free dynamic range vs intercept point (based on 2.1-kHz bandwidth, 10 dB noise figure). Maximum input signal (millivolts rms across 50 ohms) is shown at top.

For example, assuming a noise figure of 10 dB and 2.1-kHz bandwidth, the spurious-free dynamic range and maximum input signal of a receiver with an intercept point at +30 dBm are

$$DR = 2/3[30 - (-131)] = 107.3\ dB$$

$$P_{i(max)} = 1/3[2 \cdot 30 + (-131)]$$

$$= -23.7\ dBm$$

Thus, the IMD products will be well below the noise level for all input signals below −23.7 dBm (14000 μV or about S9+43 dB)! As a comparison, consider a receiver with a third-order intercept at −9.5 dBm (10 dB noise figure, 2.1-kHz bandwidth):

$$DR = 2/3[-9.5 - (-131)] = 81\ dB$$

$$P_{i(max)} = 1/3[2(-9.5) + (-131)]$$

$$= -50\ dBm$$

With this receiver the distortion products are equal to the noise level with an input signal of about 710 μV or S9+17 dB. This may represent adequate strong-signal performance if you live out in the country, but it's doubtful. If you live in an urban area, you'll have a lot of trouble digging weak signals out of the morass of IMD products which effectively raise the noise floor of the receiver.

A graph of intercept point vs dynamic range and maximum input signal is presented in **fig. 12** for a receiver with a 10 dB noise figure and 2.1-kHz bandwidth (typical for modern amateur communications receivers).

summary

Although modern amateur receivers are no longer performance limited by noisy vacuum tubes or poor selectivity, the published performance specifications have changed little since the 1940s and are still limited essentially to data on sensitivity and selectivity. Specifications on intermodulation distortion, cross modulation, desensitization and dynamic range, if they're mentioned at all, provide insufficient information for direct buyer comparison. Purchasing a new receiver under these conditions is a bit like buying a new car without knowing gas mileage or how many passengers it will carry.

The performance specifications for the high-frequency receiver shown in **table 3** leave no question as to receiver performance and are recommended as a

table 3. Specifications for a high-performance, high-frequency communications receiver provide a good format for amateur equipment manufacturers to follow. These specifications give complete reference and qualifying data and leave little question as to actual receiver performance.

Frequency range	500 kHz to 30 MHz
Tuning accuracy	±500 Hz relative to the frequency of the desired signal
Sensitivity	CW and ssb: 0.5 μV for 10 dB S/N ratio in a 2.4-kHz bandwidth (11 dB noise figure)
I-f selectivity	2.1 kHz at −6 dB, 4.2 kHz at −60 dB (2.0 shape factor)
Intermodulation products	Out of band: With two 20 mV signals separated and removed from the desired signal by not less than 25 kHz the third-order IMD products are not less than 90 dB below either of the interfering signals. Intercept point = + 24 dBm In band: Two in-band signals of 20 mV will produce third-order IMD products not greater than −50 dBm
Dynamic range	102 dB
Cross modulation	With a desired signal greater than 100 μV in a 2.4-kHz bandwidth, an unwanted signal, 30% modulated, removed not less than 25 kHz, must be greater than 175 mV to produce an output 30 dB below the output produced by the desired signal.
Compression	With a desired signal of 500 μv, an unwanted signal more than 25 kHz removed must be greater than 300 mV to reduce the output by 3 dB
Spurious response	External signals 25 kHz or more removed from the desired signal must be at least 85 dB above the level of the desired signal to produce an equivalent output Internal spurious signals are not more than 3 dB above the noise level measured in a 2.1-kHz bandwidth
AGC range	An increase in input of 110 dB above 1 μV will produce an output change of less than 6 dB

guide for receiver manufacturers to follow in the future. Well informed amateurs should demand nothing less.

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ham radio

high dynamic range receiver input stages

A new look at the input stages of high-frequency communications receivers including some circuits for improving the IMD performance over that of any present amateur equipment

Ulrich L. Rohde, DJ2LR, 425 Fairfield Avenue, Ridgewood, New Jersey 07450

Various efforts have been made in the past to build high-frequency communications receivers with wide dynamic range to combat the effects of cross-modulation and blocking. With the high level of galactic and man-made noise which is present up to about 30 MHz, receiver designers generally agree that a receiver with a 10 dB noise figure is more than adequate for high-frequency communications. This is equivalent to a sensitivity of 0.2 μV for 10 dB signal-plus-noise to noise (50-ohm input impedance, 2-kHz bandwidth). In most cases, on the high-frequency bands, the man-made noise which is picked up by the antenna is greater than 0.07 μV, so the extraordinary sensitivity of a receiver with a noise figure lower than 10 dB is no great advantage. While this technique was initiated ten years ago by my company, Rohde & Schwarz, in Munich, amateur equipment designers have only recently started thinking along the same lines.

A noise figure of slightly less or equal to 10 dB can be readily achieved by using one of the new hot-carrier diode double-balanced mixers, such as the Minilabs SRA3H or SRA1H, which is

followed by a low-noise i-f amplifier ahead of a crystal filter. In practice this means that no rf amplifier stage is required. In vacuum-tube receivers one of the primary applications for the rf input stage was to provide agc but this can now be obtained (with negligible losses) with a PIN diode attenuator.[1,2]

The first amateur equipment to use this new design technique is the Atlas 180 and its later models. There are some indications that this circuit is derived from the SouthCom SC130 Man-Pack Transceiver* and the AN/PRC-104 of Hughes Aircraft, which both use the same technique. The main disadvantage of this rf input circuit is that it is practically impossible to suppress feedthrough energy from the oscillator to the antenna to 15 μV or less, a requirement of European regulatory agencies.

Some military and systems-oriented receivers still require noise figures of 6 dB or less because they are used with inefficient antennas (whip antennas with loading coils for marine, mobile or portable use, for example). Since it is difficult for the commercial communications equipment manufacturer to foresee whether the receiver will be used with a good antenna or a poor one, coupled with the requirement for low oscillator feedthrough, in many cases rf amplifier stages are still required.

Recently the German amateur journal, *cq-DL*, published an extensive test report on the Atlas 180[3] which showed fairly poor dynamic range with respect to what would be expected from a receiver with a double-balanced mixer input and high-power i-f stage before the crystal filter. As pointed out in a previous article,[4] to obtain the specified performance every mixer must see a purely resistive load at *both* the i-f and image frequency. An analysis of the SouthCom and Atlas circuit reveals a tuned circuit between the double-balanced mixer and the i-f stage. Since the mixer is not properly terminated at the image frequencies, this results in a loss of at least 15 dBm for the mixer's third-order intercept point, enhancing intermodulation distortion products.

table 1. Third-order intercept point of Minilabs SRA1H high-level double-balanced mixer with various terminations.

termination	intercept point
50-ohm resistance	30 dBm
Narrow-band resonant circuit	8 dBm
Heavily damped resonant circuit	17 dBm
Elliptical bandpass circuit	21 dBm
Amplifier (Z_i = 50 ohms ±10°, 1-80 MHz)	23 dBm
Power fet (Z_i = 50 ohms ±5°, 1-108 MHz)	30 dBm

mixer circuits

A recent survey published in Germany[5] showed the results of a series of tests regarding the creation of unwanted spurious in double-balanced mixers as a function of termination. All tests were run with a high-level Minilabs SRA1H mixer which requires a local-oscillator input of 23 dBm (200 milliwatts). The results of these tests are listed in **table 1**.

Although the Minilabs SRA1H is capable of providing a third-order intercept point of 30 dBm, **table 1** indicates

*Atlas Radio is licensed by SouthCom International to use the circuits developed by SouthCom for military and commercial communications equipment. **Editor.**

that you can lose as much as 22 dB of dynamic range because the mixer is not properly terminated. Therefore, to obtain maximum performance, it is essential to build an input stage which has the required input impedance from dc be adjusted to set the dc bias so the input impedance is exactly 50 ohms.[5] This circuit has been designed for optimum dynamic range regardless of oscillator feedthrough and does not show any input selectivity which must, of course, be

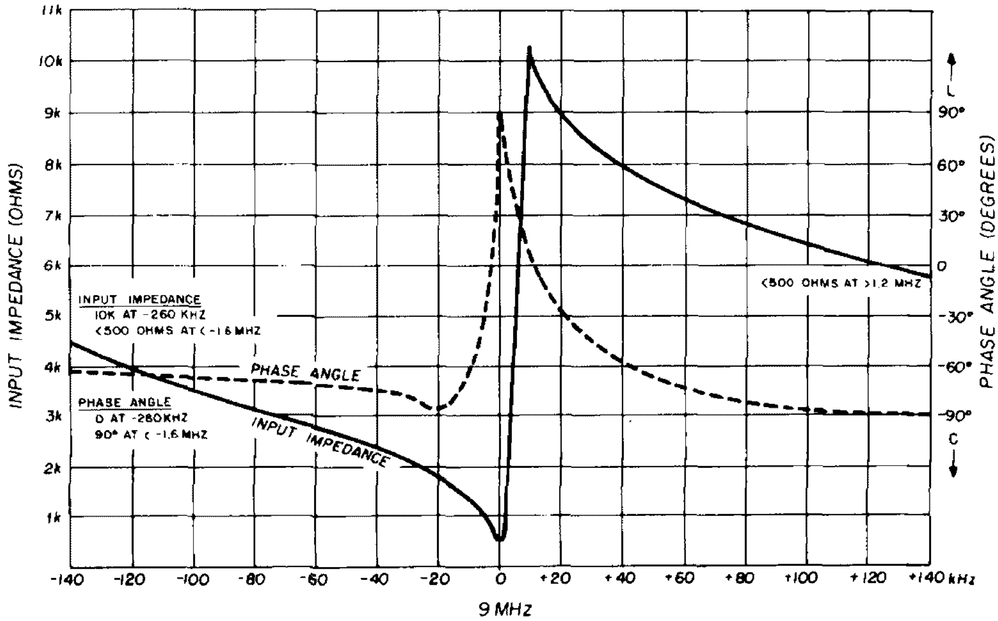

fig. 1. Input impedance and phase angle of the 9-MHz KVG XF9B crystal filter when terminated in 560 ohms in parallel with 33 pF.

up to more than 100 MHz. This stage must also have an intercept point of 25 dBm (the double-balanced mixer has about 5 dB loss so the mixer output is 25 dBm at 30 dBm input).

Extensive tests with various crystal filters have been conducted in the past and it was found that the crystal filter's input and output must be terminated beyond its normal frequency range of operation. This is because crystal filters, such as the KVG XF9B, at frequencies not too far from the center frequency, exhibit impedances between 5 and 10 kilohms (see fig. 1). These resonant effects would significantly reduce the dynamic range of the preamplifier if the filter was not properly terminated.

The best solution to the problem of a high dynamic range preamplifier is a high current field-effect transistor, type CP643 (Teladyne Crystalonics), in the circuit of fig. 2 where the input resistors can added. In most cases input selectivity can be provided by a 1.6-MHz highpass filter and a 31-MHz lowpass filter, so the losses there do not have to be considered.

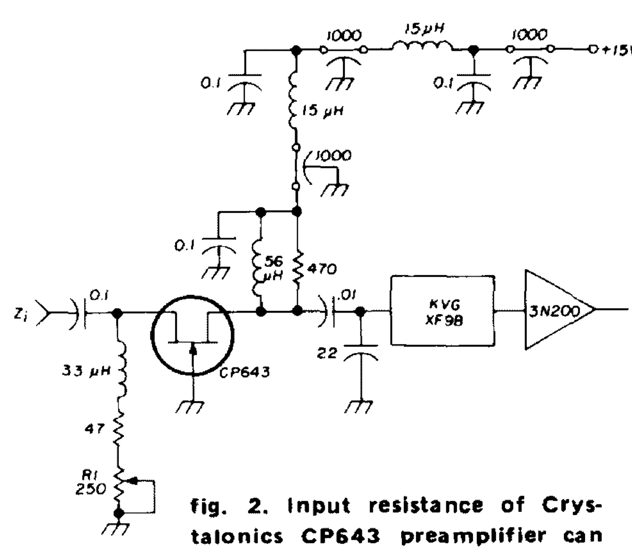

fig. 2. Input resistance of Crystalonics CP643 preamplifier can be adjusted to 50 ohms by proper setting of the 250-ohm potentiometer, R1. Current drain is 30 mA.

The complete rf input circuit, showing the CP643 high dynamic range preamplifier, double-balanced mixer, 2N5109 oscillator injection amplifier and 3N200 i-f amplifier is shown in **fig. 3**. The injection amplifier accepts about 200 mV input voltage from the local oscillator and amplifies it to the required level for the double-balanced mixer. The grounded-gate CP643 has 14 dB gain and is matched to the input of the KVG XF9B crystal filter. The 3N200 i-f amplifier has enough gain and agc action for most receiver designs.

push-pull rf amplifier with wide dynamic range

In many wideband, high dynamic range applications such as antenna distribution amplifiers, input rf amplifiers are required which combine extremely low distortion with low noise figure. In the past both voltage and current feedback have been used to counteract voltage and current distortion. The disadvantage of these circuits is that a stable input impedance can be achieved only over a relatively narrow bandwidth.[6]

Recent research has resulted in a new wideband amplifier design[2,7] which has extremely low vswr at both input and output as well as low noise figure. The

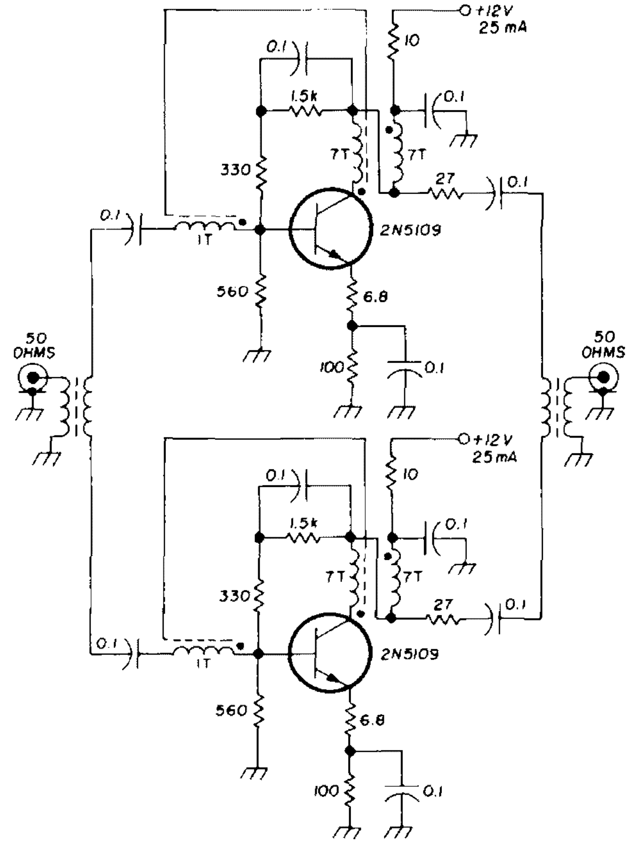

fig. 4. Vhf power transistors are used in push-pull circuit to obtain wide dynamic range shown in fig. 5. Transformers are trifilar wound on Indiana General F625-9-TC9 toroid cores.

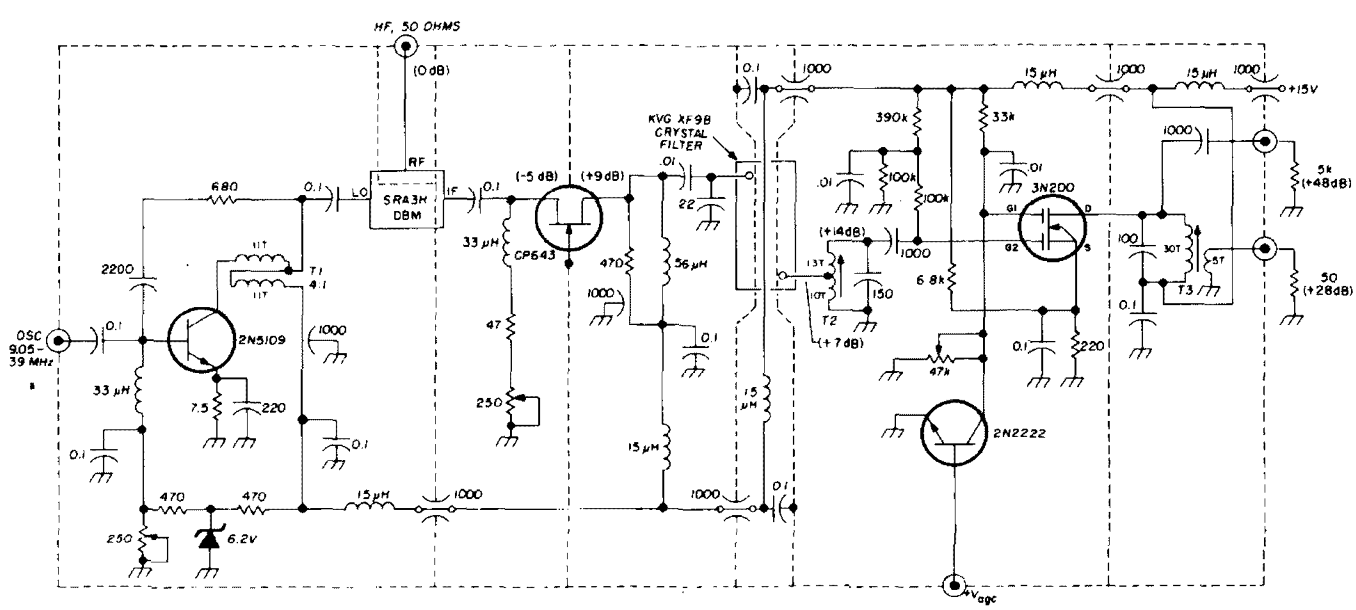

fig. 3. Double-balanced mixer circuit which has a third-order intercept point at +30 dBm. Oscillator requirement is −1 to +2 dBm (200 to 280 mV across 50 ohms). Agc range is greater than 50 dB. Levels shown in parenthesis are for zero dBm (224 mV) at input and zero agc voltage.

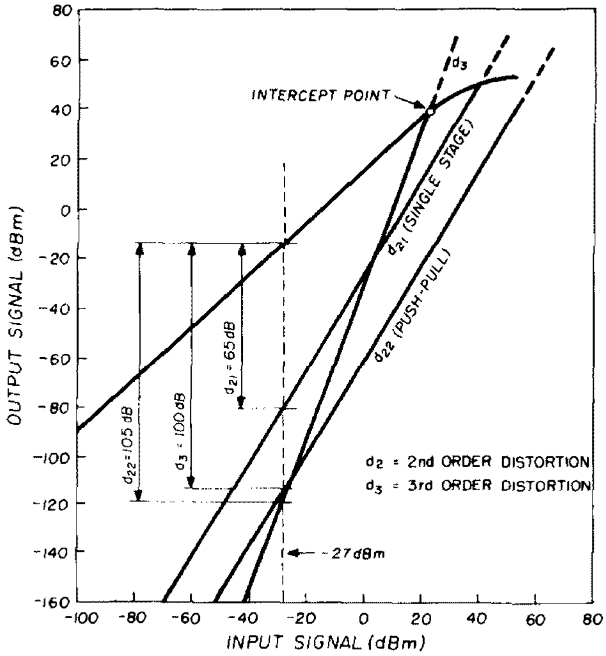

fig. 5. Performance of the push-pull rf amplifier shown in fig. 4. With an input of −27 dBm (two-tone signal, 20 mV each), gain is 12 dB, third-order distortion products are down 100 dB and second-order IMD is down 105 dB. Third-order intercept point occurs at an input of about 22 dBm.

second-order intermodulation products can be suppressed nearly 40 dB (over a single stage) by the push-pull arrangement shown in **fig. 4**. Two linear vhf power transistors are used in the circuit, and depending upon the large-signal handling requirements, either the 2N5109 (RCA) or BFR95 (Amperex) may be used. Both of these devices have an F_T of 1600 MHz.

This circuit provides about 11 dB gain and exhibits exceptional freedom from second- and third-order intermodulation products, as plotted in **fig. 5**. The third-order intercept point occurs at an input of about 22 dBm. Three types of feedback are used: *current* feedback through the unbypassed 6.8-ohm emitter resistor, *voltage* feedback through the unbypassed 330-ohm base-to-collector resistor, and *transformer* feedback through a third winding on the wideband transformer to stabilize the input and output impedance. A mathematical analysis of this circuit is presented for interested readers in the appendix.

table 2. Measured third-order intercept point of several commercial high-frequency receivers.

receiver	intercept point
Yaesu FT101	−21.5 dBm
Ten-Tec Argonaut	−19.5 dBm
Collins KWM2/S-line	−10.0 dBm
Signal 1 CX7	−5.0 dBm
Collins R390A	−4.5 dBm
Atlas 180/210	3.0 dBm
Collins 65S1	13.0 dBm
Racal RA1772	28.0 dBm
Martin rf front end (fig. 3)	30.0 dBm

summary

With very little effort the third-order intercept point of high-frequency receiver input stages can be increased far beyond the values of any commercial equipment now on the market. **Table 2** shows the third-order intercept point of several popular receivers presently being used by amateurs.

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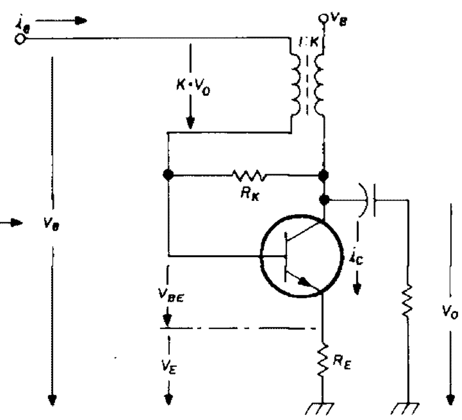
appendix

Since voltage and current feedback may result in input and output impedances which may not suit the design requirements, a wideband toroidal transformer with a high-permeability core (such as Indiana General, F625-9-TC9) can be used to arbitrarily set these impedances. The schematic below



shows an amplifier using a transformer with voltage and current feedback. The input voltage, v_e, input current, i_e, and input impedance, Z_e, are given by the following equations

$$v_e = (k \cdot v_o) + v_{be} + v_e$$

$$i_e = (v_{be} \cdot Y_{11}) + v_{cb}/R_k$$

$$Z_e = \frac{(k \cdot v_o) + v_{be} + v_e}{(v_{be} \cdot Y_{11}) + v_{cb}/R_k}$$

So long as the operating frequency is well below f_T, emitter and collector current are the same. Therefore, the input impedance of the stage will be

$$Z_e = R_k \left(\frac{k + A}{1 - A}\right)$$

where $$A = \frac{R_E (R_k + R_L)}{(R_E + R_k) R_L}$$

and the required value for the voltage feedback resistor, R_k, is given by

$$R_k = Z_e \left(\frac{1 - A}{k + A}\right)$$

the amplification of the stage is given by

$$A = \frac{1}{k + \dfrac{R_E (1 + R_k/R_L)}{R_k + R_E}}$$

Since it is advantageous to have a 50-ohm input impedance, and the output impedance of the stage is approximately 150 ohms, the collector winding is split to build a 4:1 transformer. Under these conditions the input impedance, Z_e, output impedance, Z_o, and voltage gain, A, are given by

$$Z_e \approx (k \cdot R_k) + \frac{R_E}{2}$$

$$Z_o \approx \frac{R_E}{2k}$$

$$A = \frac{1}{k\left(1 + \dfrac{R_E}{2k \cdot Z_L}\right)}$$

The constant k, which determines the turns ratio between the base and collector coil (1:7:7 in **fig. 5**), can only be an integer. To obtain optimum performance in many cases, therefore, one of the values may have to be a compromise. In the circuit of **fig. 5** the output impedance was only 23 ohms so a 27 ohm resistor had to be placed in series to obtain the desired 50-ohm output impedance.

ham radio

capacitance meter

Dear HR:

I have receiver a number of queries regarding the programmable unijunction transistor used in the capacitance meter described in the April, 1975, issue of *ham radio*. The full part number of this device, which is manufactured by Texas Instruments, is A7T6028 (because of space limitations, the package is labeled AT6028). The 2N6027, 2N6028 and 2N6118 are similar.

Although I have not tried the Motorola HEP S9001 programmable unijunction in the circuit, I have letters from K6MYA and WA3IFQ who say they had excellent results with this device. Another possibility is the package of programmable unijunctions available from Radio Shack (part number 276-119).

The two series-connected 0.005 μF capacitors in the circuit may be replaced by a single 0.0025 μF capacitor. This apparently has caused some confusion.

Courtney Hall, WA5SNZ
Dallas, Texas

solid-state communications receiver

A five-band amateur-band receiver which features an active balanced mixer front end for improved intermodulation performance

Piero Moroni, I5TDJ, Cosseria 10, Florence, Italy 50129

Radio amateurs in Italy live in a special purgatory. The American ham magazines show us the beautiful equipment which is available, but customs duties and air shipment add about 100 per cent to the list price. This is probably one reason why we have a larger percentage of homebrew equipment here in Italy.

My first solid-state receiver[1] had been working fairly nicely, but I wanted something with a little better dynamic range, especially on 7 MHz as broadcast stations on this band are a little more bothersome here in Europe. A receiver with better cross modulation and blocking characteristics would help. The receiver described here seems to have achieved this objective, although I am an inveterate experimenter and will no doubt make a few changes. My friend Luciano, I5FLN, who duplicated my circuitry and packaged it very nicely, calls his copy the 5-Band DXCC reveiver (see photo above). He has worked over 150 countries on 7-MHz ssb. My re-

ceiver isn't as pretty — I was too impatient to hear the results!

The receiver uses a mixture of U.S. and Philips transistors which I accumulated, but I have suggested alternate types which are readily available in the States. Enterprising amateurs should have no trouble substituting equivalents from the myriad of types available. Performance characteristics of the receiver are listed in **table 1**.

circuit description

A block diagram of the receiver is shown in **fig. 1**. The signal at the antenna first goes through bandpass filters, one for each band. Each filter consists of three capacitively-coupled tuned circuits which use high-Q toroids to reduce losses and suppress out-of-band signals.

This article was translated for *ham radio* by Josef Darmento, W4SXK. He reports that Piero is an engineer with a major electronic manufacturer in Florence, and that his ham shack is full of "auto constructed" (homebrew) equipment including some test instruments. Joe visited Piero in the fall of 1974 and had an opportunity to use the receiver described in this article. As Joe tells it, it was pure pleasure, but "how much was due to the receiver and how much was due to a different environment would be very difficult to evaluate."

A balanced mixer following the filter insures that local oscillator noise is down some 30 dB. The vfo covers 5.0 to 5.5 MHz and effects the first conversion on 80 and 20 meters; for reception on 40, 15 and 10 meters the vfo is heterodyned with crystal oscillators for mixer injection.

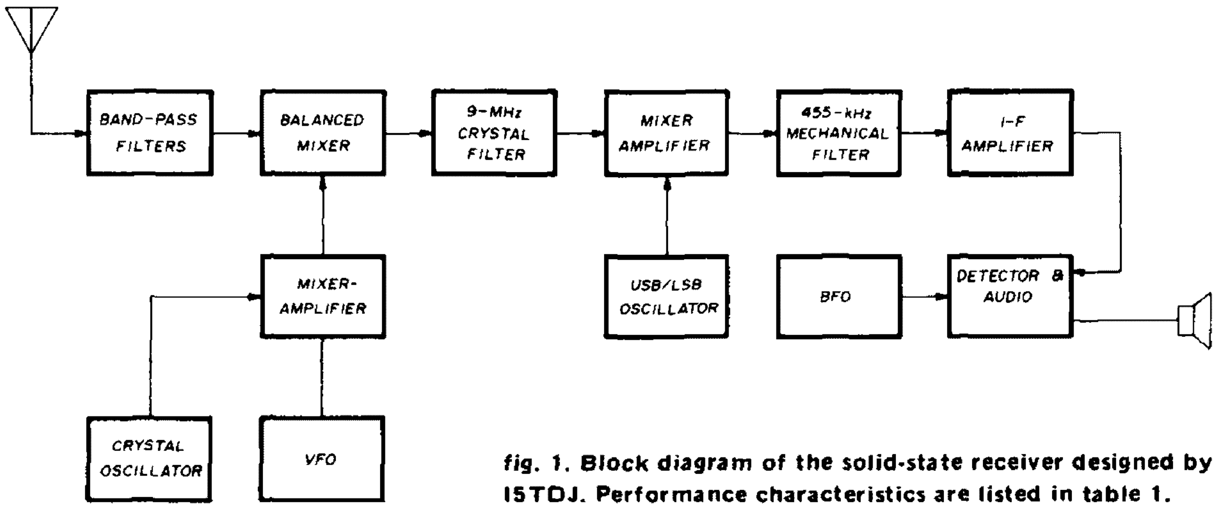
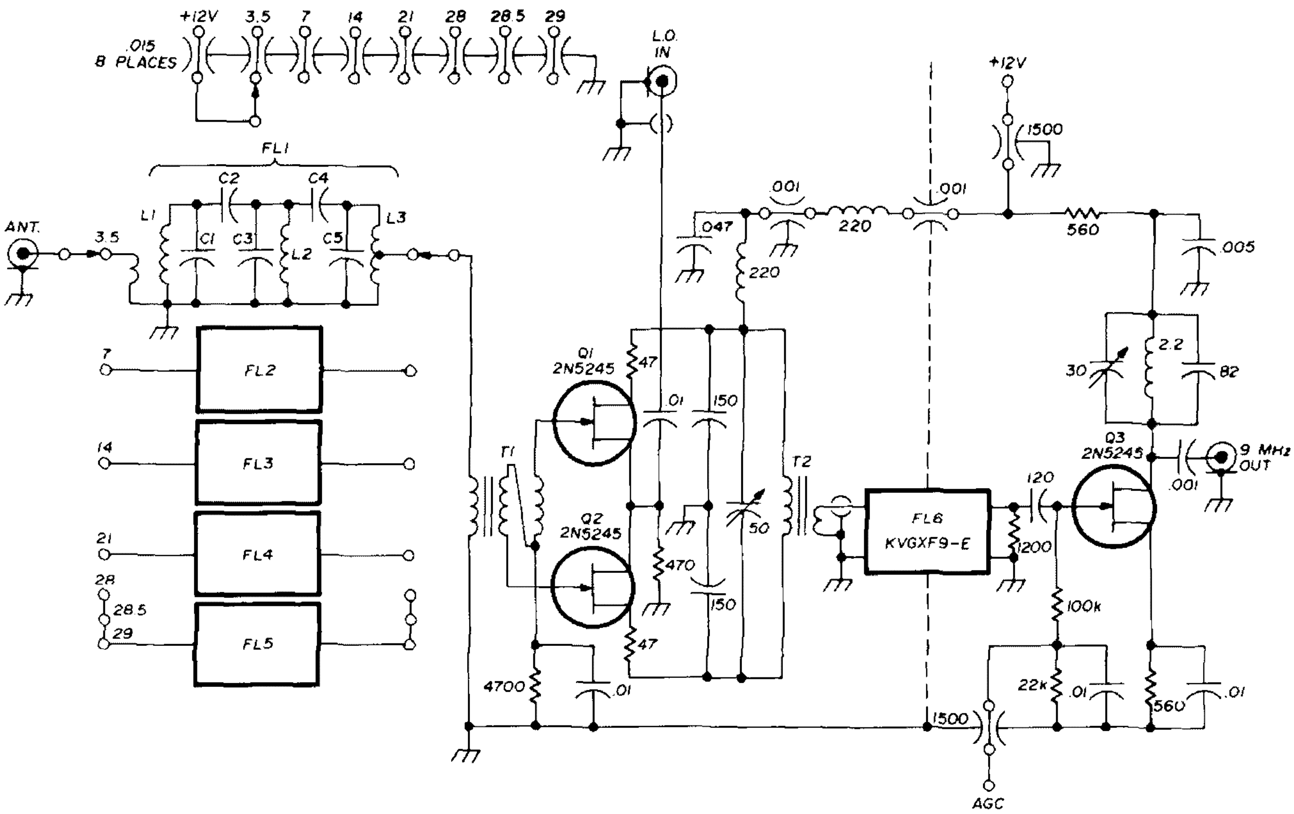


fig. 1. Block diagram of the solid-state receiver designed by I5TDJ. Performance characteristics are listed in table 1.

The output of the balanced mixer is applied to a 9-MHz crystal filter, a KVG XF9E. This is an 8-pole filter with a 6-dB bandwidth of 2.4 kHz, a shape factor of 1.8 and stopband attenuation

table 1. Performance of the I5TDJ communications receiver.

Frequency coverage	80 through 10 meters
Sensitivity	0.25 µV for 10 dB signal-plus-noise-to noise ratio
AGC range	110 dB (3 dB increase in audio for antenna signal increasing from 1 µV to 300 mV)
Desensitization	70 mV (−13 dBm) signal ±100 kHz from desired 0.25 µV (−117 dBm) signal lowers signal-to-noise ratio from 10 dB to 3 dB
Intermodulation	With 5 mV (−33 dBm) signals ±50 kHz from desired 0.2 µV (−121 dBm) signal, distortion products are 88 dB below the interfering signals.



		L1	L2	L3	C1	C2	C3	C4	C5
FL1	3.5-4.0 MHz	14.2 μH, 59 turns no. 28 (0.03mm); 12-turn link	Same as L1, no link	Same as L1, tap 42 turns from ground	130	15	100	15	115
FL2	7.0-7.3 MHz	2.3 μH, 23 turns no. 24 (0.5mm); 5-turn link	Same as L1, no link	Same as L1, tape 17 turns from ground	200	8.2	200	8.2	200
FL3	14.0-14.5 MHz	0.97 μH, 16 turns no. 22 (0.6mm); 3-turn link	Same as L1, no link	Same as L1, tap 10 turns from ground	120	3.9	120	3.9	120
FL4	21.0-21.5 MHz	0.48 μH, 10 turns no. 22 (0.6mm); 2-turn link	Same as L1, no link	Same as L1, tap 7 turns from ground	120	3.3	120	3.3	120
FL5	28.0-29.7 MHz	0.48 μH, 10 turns no. 22 (0.6mm); 2-turn link	Same as L1, no link	Same as L1, tap 7 turns from ground	60	2.2	60	2.2	60

Note: All inductors for FL1 and FL2 are wound on Amidon T50-6 toroid cores; inductors for FL3 FL4 and FL5 are wound on Amidon T50-10 toroid cores.

T1 10 turns no. 32 (0.2mm), trifilar wound on Amidon T50-6 toroid core

T2 Primary is 22 turns no. 28 (0.3mm), evenly distributed on Amidon T50-6 toroid core; secondary is 7 turns no. 28 (0.3mm) over center of primary winding

fig. 2. Input filters, balanced mixer and 9-MHz amplifier. Filter capacitor values are in pF. Construction of this module is shown in fig. 3.

greater than 100 dB. A 9-MHz fet amplifier follows and a second fet mixer converts the 9-MHz signal to 455 kHz for application to a 2.7 kHz wide Collins mechanical filter. Crystal oscillators at 8545 and 9455 kHz provide sideband selection and are selected automatically with the bandswitch in accordance with current convention. A 455-kHz i-f amplifier, balanced diode detector, agc and an audio IC complete the picture. All circuitry derives power from an adjust-

able, regulated +12 volt power supply. Total current drain is about 375 mA.

construction

As indicated in the previous description, the receiver contains nothing new or startling. However, it is in the construction that careful attention to detail pays off. The receiver was built in modules, each module housed in a 177x145x49mm (7x5.7x2 inch) high aluminum box, a commercially available size. Subdivisions within the boxes contain the functional sections, each on a piece of one-sided copper fiberglass circuit board. This was done to facilitate system modification. Modular construction also contributes to greater freedom from spurious products. All power leads enter the modules through 1500-pF feedthrough capacitors. Signal interconnections are by phone jacks.

mixer module

The input filters and the balanced mixer are on four small boards grouped around the bandswitch. The 9-MHz amplifier is further isolated in its own smaller box (see fig. 3). Bandpass filters F1 and F2 are on one board, F3 and F4 are on a second board, F5 is on a third and the balanced mixer is on a fourth. The objective here was to realize the maximum attenuation capability of the filters.

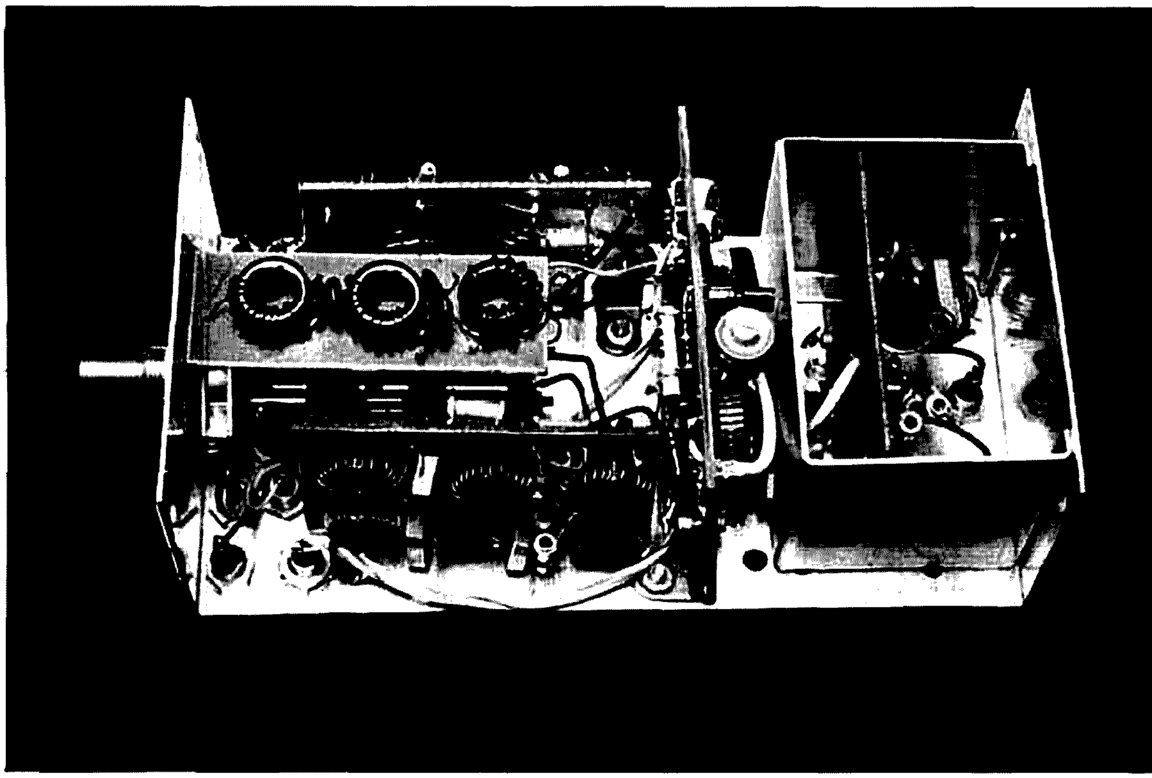


fig. 3. Mixer module showing the input filters (left) and 9-MHz amplifier (separate enclosure at right). The balanced mixer is on the vertical board in the center.

Transformer T1 is trifilar wound with a single primary and two secondaries in series. Transistors Q1 and Q2 should be selected for equal *drain* current with the gate at zero volts. The small box with the 9-MHz amplifier is completely shielded and isolated with a bottom and top cover of tinned copper sheet and is fastened with epoxy to the larger aluminum box. Only thus was I able to realize the maximum attenuation capabilities of the KVG filter; the filter itself is mounted externally. The gain of Q3 is agc controlled. Note that the leads carrying band-change informa-

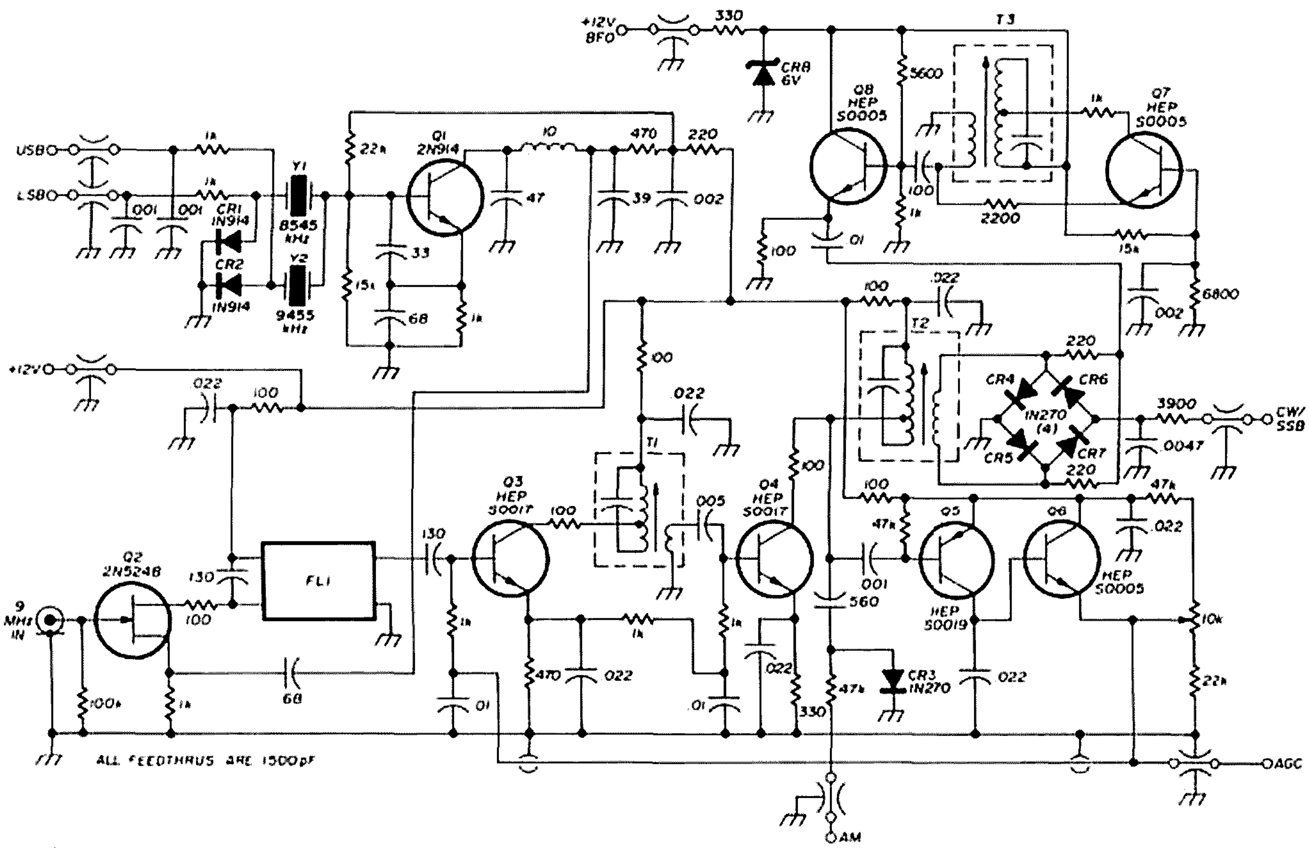


fig. 4. Circuit of the 455-kHz i-f module. The mechanical filter, FL1, is a Collins F455Z7. T1, T2 and T3 are miniature 455-kHz i-f transformers (see text). Module construction is shown in fig. 5.

tion enter the module through feed-through capacitors.

455 kHz module

The 455-kHz amplifier is on a single board (see figs. 4 and 5). The Collins filter provides good isolation between input and output so no elaborate precautions are necessary. I mounted the transistors in sockets for easy substitution. The four diodes used in the product detector, CR4-CR7, are Phillips AAZ15 (1N270) selected for equal forward resistance. The BF175 transistors,

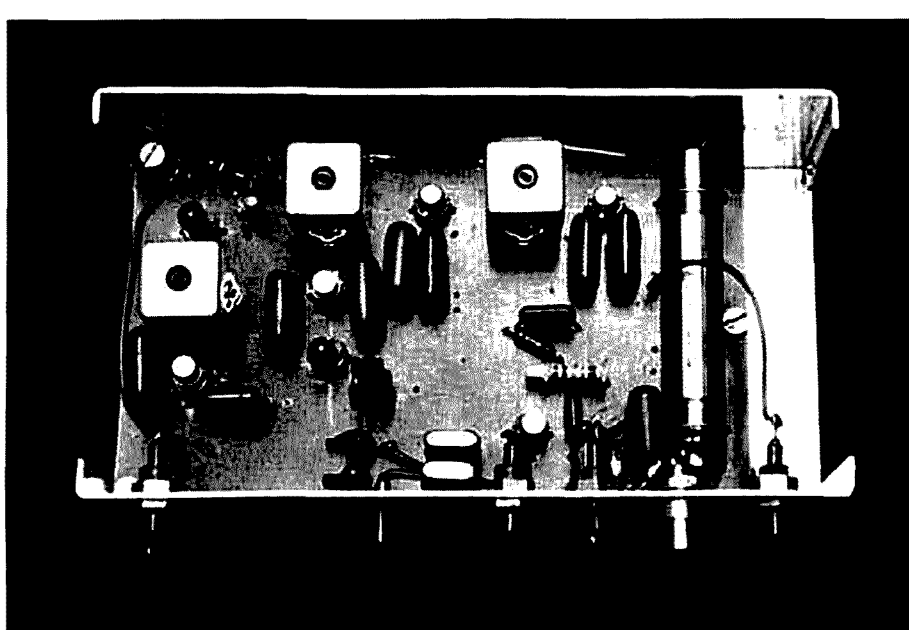


fig. 5. The 455-kHz i-f module. The Collins mechanical filter is at right.

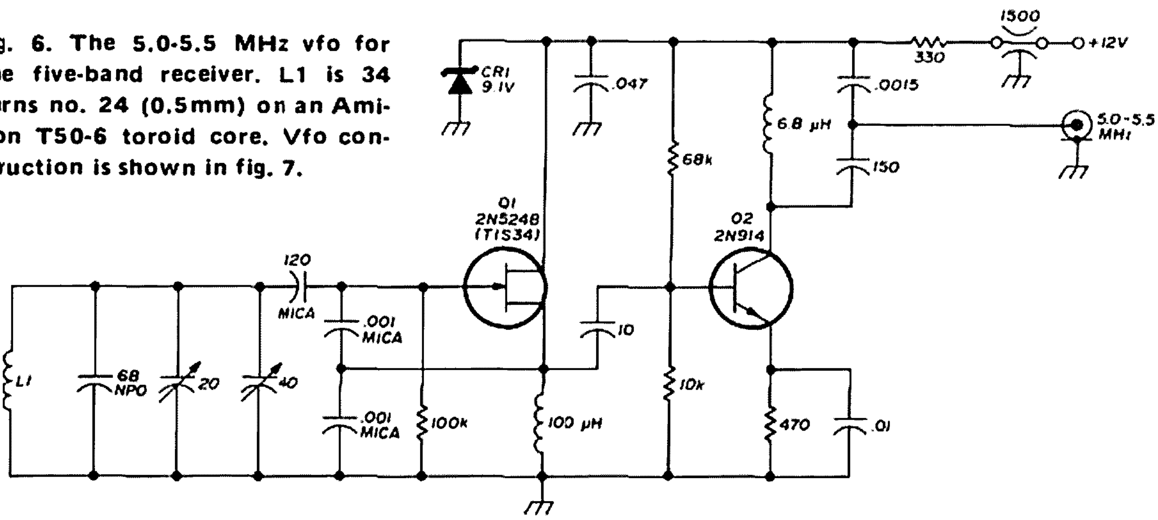


fig. 6. The 5.0-5.5 MHz vfo for the five-band receiver. L1 is 34 turns no. 24 (0.5mm) on an Amidon T50-6 toroid core. Vfo construction is shown in fig. 7.

Q3 and Q4, are rf types with gain which diminishes with increases in collector current. The Fairchild 2N4134 or HEP S0017 are good substitutes; these types may also be substituted for the BC209 and EC209. Note that transistor Q5, a BC154, is a pnp type; a HEP S0019 may be used instead.

Transformers T1, T2 and T3 are slug-tuned Siemens transformers – actually a bit large (25mm square) for this application. The J. W. Miller series 2031 i-f transformers only 0.5 inch square (12.5mm) might be more fitting if you have small fingers. The bfo may be crystal controlled if desired; its frequency, of course, will depend on the filter used in the receiver.

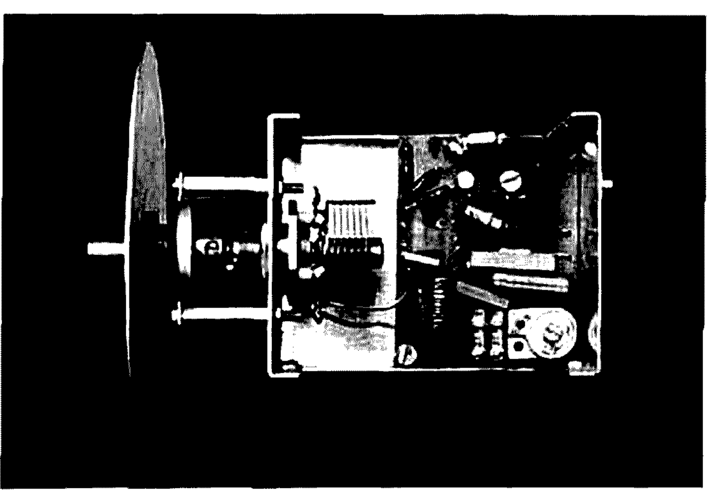


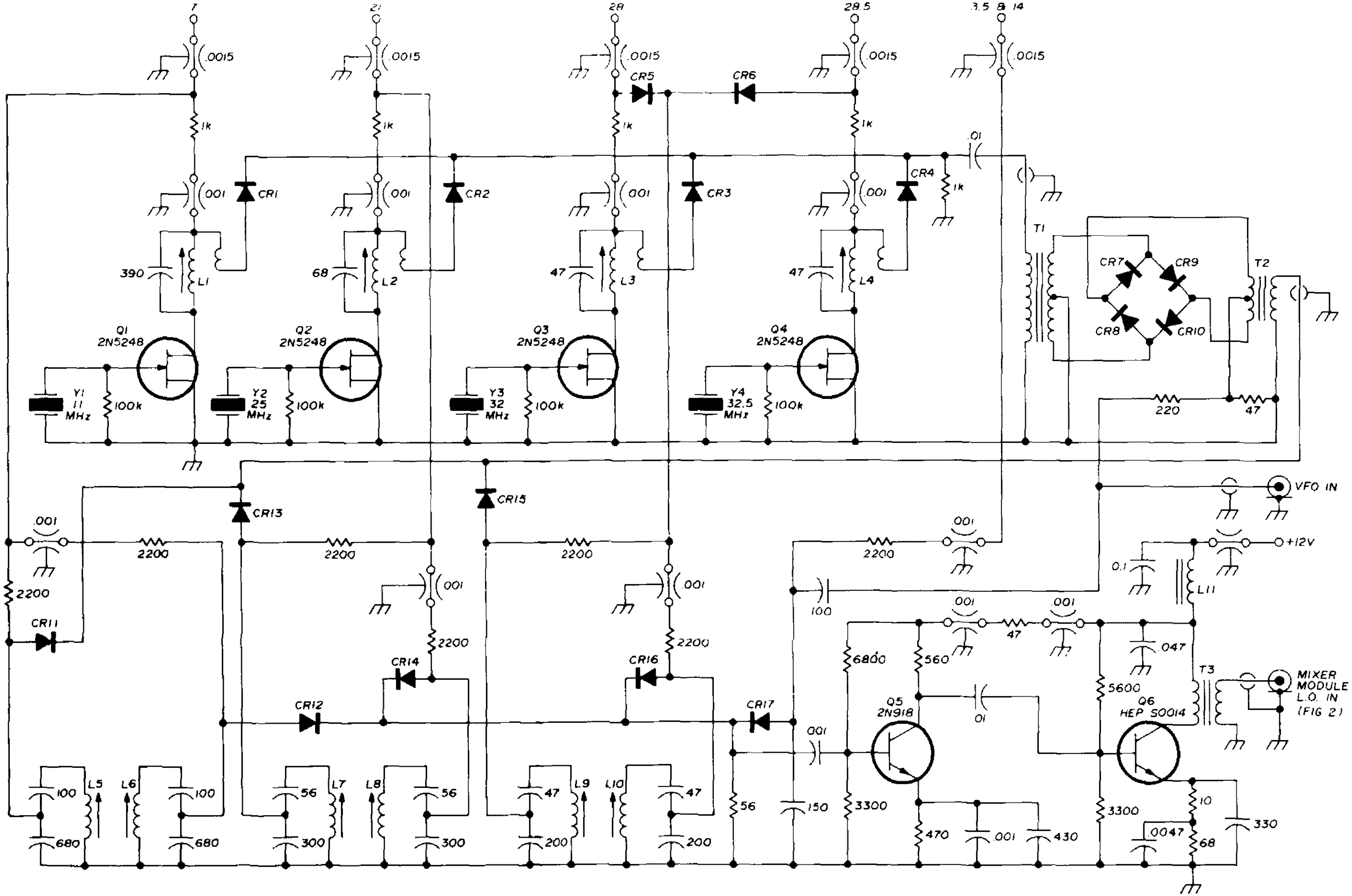
fig. 7. Construction of the 5.0-5.5 MHz vfo. A 5:1 planetary drive provides smooth tuning.

vfo module

The vfo is in a box which I strengthened by various methods to achieve mechanical rigidity (see figs. 6 and 7). The components are on a small fiberglass board except for the small 40 pF variable capacitor. The 68 pF NPO affords temperature compensation and the 20 pF trimmer sets the band center. A 5:1 planetary drive gives nice easy tuning. A digital readout would have been even nicer, but would use a lot more power. Toroid L1 is installed edgewise on the board by its leads and a small dab of epoxy. I5FLN used a Drake vfo which makes an excellent substitution with less work.

vfo converter

The circuitry and construction of the vfo converter module are shown in figs. 8 and 9. The 5.0-5.0 MHz vfo signal at the vfo input goes directly to amplifiers Q5 and Q6 when the bandswitch is on 3.5 or 14 MHz. When applied to the balanced mixer the product is at 9 MHz. When the bandswitch is on 7, 21 or 28 MHz the vfo signal is mixed with the output of an fet crystal oscillator and filtered before being applied to Q5 and Q6. Two crystal oscillators at 28 and 28.5 MHz cover important segments of the 10-meter band. The filters suppress undesired products from the diode mixer, CR7-CR10. These diodes should


7
21
28
28.5
3.5 & 14
.0015
1k
.001
.01
CR1
CR2
CR3
CR4
CR5
CR6
390
68
47
L1
L2
L3
L4
Q1 2N5248
Q2 2N5248
Q3 2N5248
Q4 2N5248
Y1 11 MHz
Y2 25 MHz
Y3 32 MHz
Y4 32.5 MHz
100k
T1
T2
CR7
CR8
CR9
CR10
220
47
CR13
CR15
VFO IN
2200
CR11
CR12
CR14
CR16
CR17
+12V
0.1
L11
100
.047
6800
560
5600
T3
MIXER MODULE L.O. IN (FIG 2)
Q5 2N918
Q6 HEP S0014
.01
L5
L6
L7
L8
L9
L10
680
56
300
200
150
3300
470
430
.0047
10
68
330

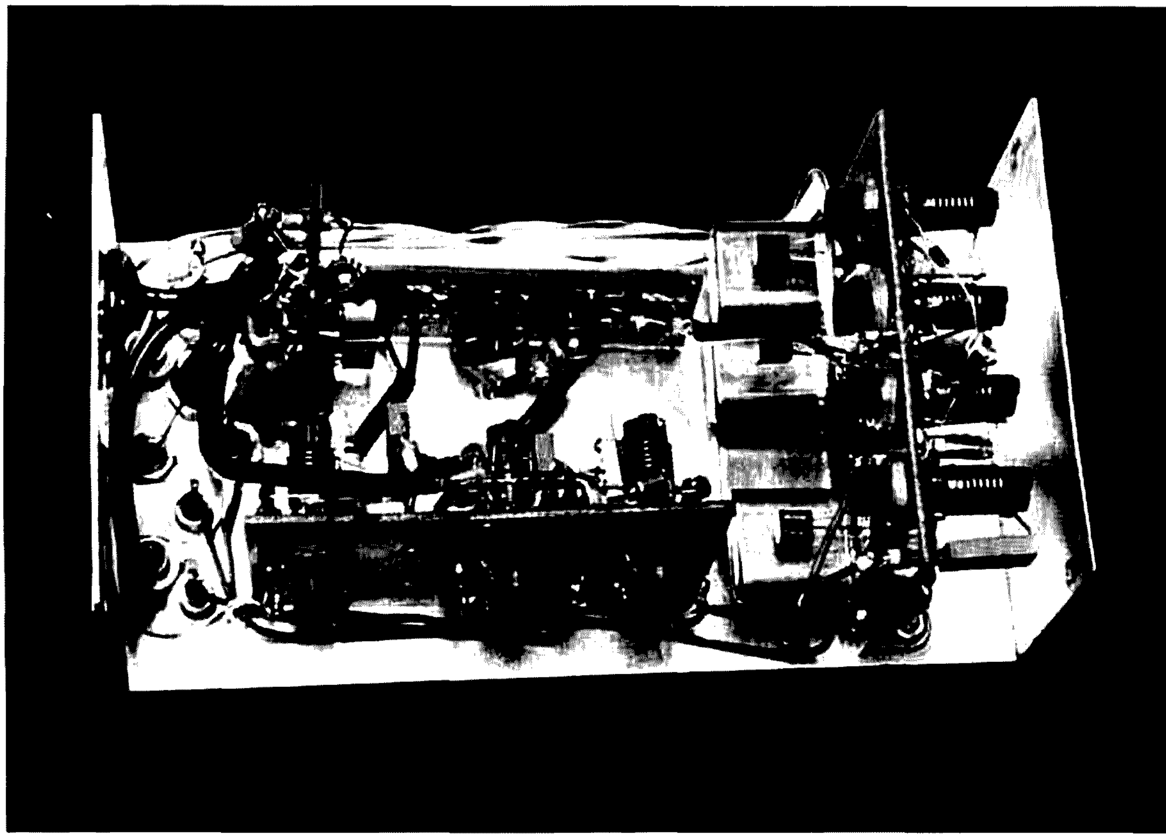


fig. 9. The vfo converter module. Filter pairs L5-L6, L7-L8 and L9-L10 are on the board in the foreground. Crystal oscillators are at the far right.

fig. 8. Vfo converter (left). All switching is accomplished with semiconductor diodes. Crystals are parallel resonant with a 32 pF load; Y2, Y3 and Y4 are third-overtone types. Construction of the module is shown in fig. 9.

CR1-CR6	1N914 or equivalent
CR7-CR10	Selected 1N270 diodes (see text)
CR11-CR17	1N914 or equivalent
L1-L4	0.6 μH (10 turns no. 22 (0.6mm) enamelled on 3/8" (9mm) diameter slug-tuned forms. Link is 3 turns no. 22 (0.6mm)
L5,L6	1.2 μH. 20 turns no. 28 (0.3mm) on 3/4" (9mm) diameter slug-tuned forms
L7-L10	0.6 μH (same as L1 - L4)
T1, T2	10 turns no. 32 (0.2mm), trifilar wound on Amidon T50-6 toroid core
T3	10 turns no. 32 (0.2mm), trifilar wound on Amidon T50-6 toroid core. Collector winding has two windings in series to give 2:1 ratio

be carefully selected for equal (±20 mV) voltage drop at varied values of current, say 0.75, 2, 10 and 20 mA. This is a must if you want oscillator attenuation of 30 to 40 dB.

The four fet oscillators are on one board and are energized by +12 volts from the bandswitch; diodes CR1-CR4 select the output. Drain coils L1, L2, L3 and L4 are wound on slug-tuned forms. The wideband trifilar transformers, T1 and T2, are on a second board with the diode mixer. The next board has the three filter pairs, L5-L6, L7-L8, and L9-L10. Each pair is inductively coupled by mounting the two coils about one diameter (9mm or 0.4 inch) apart. Input and output for the filters is via CR11 through CR17.

The fourth board has wideband amplifiers Q5 and Q6. The BFY63 transistor is a high-gain rf type which may be replaced by a HEP S0014. Since my

design called for amplification from 5 to 39 MHz I used as short leads as possible and ceramic capacitors. L11 is a small rf choke and consists of 20 turns or so on a 5mm (0.2 inch) ferrite rod. T3 is wide-band and trifilar wound.

audio and power

Fig. 10 shows the rest of the receiver and fig. 11 the interconnecting diagram.

alignment

After installation in their respective modules, I initially set most of my toroids with a grid dipper coupled with a one-turn loop, and a receiver for frequency accuracy. The job is simplified if you have a frequency counter. A slight loosening or tightening of the wire will bring the toroid to desired resonance. For a visual indication I used a vtvm

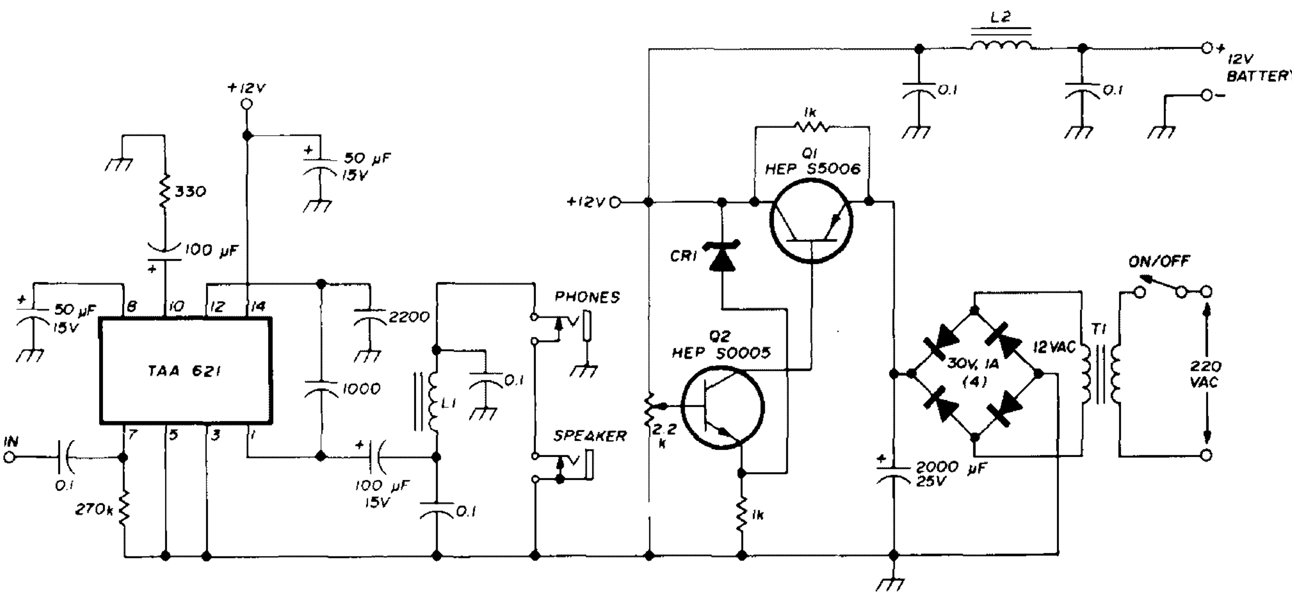


fig. 10. Audio amplifier and regulated power supply. A three-terminal 12-volt IC regulator such as the 7812 may be used, if desired. L1 and L2 are each 20 turns no. 18 (1mm) on a 5mm (0.2 inch) ferrite rod.

I did not bother enclosing the audio section or power supply. Since I am not a high-fidelity addict I used an audio IC which gives adequate volume for earphones or speaker. The TAA621 IC provides about 1 watt output with a 12-volt supply; an HEP C6093C or MC1454G may be substituted. L1 is a 5mm (0.2 inch) ferrite rod with 20 turns number-18 (1mm) wire; it and the two capacitors inhibit rectification of strong out-of-band rf signals (L2 in the power supply is the same).

I rewound an old transformer to give me about 12 volts and added a simple regulator. If you have a three-terminal voltage-regulator IC such as the MC7812, by all means use it.

with an rf probe. For the input filters I put the probe on the gate of Q1 and set the ±1 dB points of the filters as follows:

FL1	3.5 - 3.8 MHz
FL2	7.0 - 7.3 MHz
FL3	13.9 - 14.5 MHz
FL4	21.0 - 21.5 MHz
FL5	28.0 - 29.5 MHz

For inductances L1, L2, L3, L4 and L5 (fig. 8) I coupled the receiver to T1 and energized the respective oscillator. The probe on the hot side of T1 indicated about 0.5 volt for each oscillator as each coil was adjusted to resonance. Adjustment of the three filters is more easily done with vfo injection; alternately, a signal generator and rf voltmeter may be

used to set their −3 dB response as follows:

L5-L6	16.0 - 16.3 MHz
L7-L8	30.0 - 30.5 MHz
L9-L10	37.0 - 38.0 MHz

I used a calibrated receiver to set the center of the vfo excursion with the 20 pF trimmer; rf output was 0.6 volt. I connected the vfo to the converter input connector and temporarily installed receiver S+N/N at 10 dB. I then introduced two more signals to the antenna, 5 mV each, at 28.75 and 28.8 MHz. The receiver output remained at 10 dB, i.e., distortion products were 88 dB below the two interfering signals.

For the blocking check I used one signal generator as before, 0.28 μV (−117 dBm) on 28.7 MHz, then introduced a second generator and adjusted

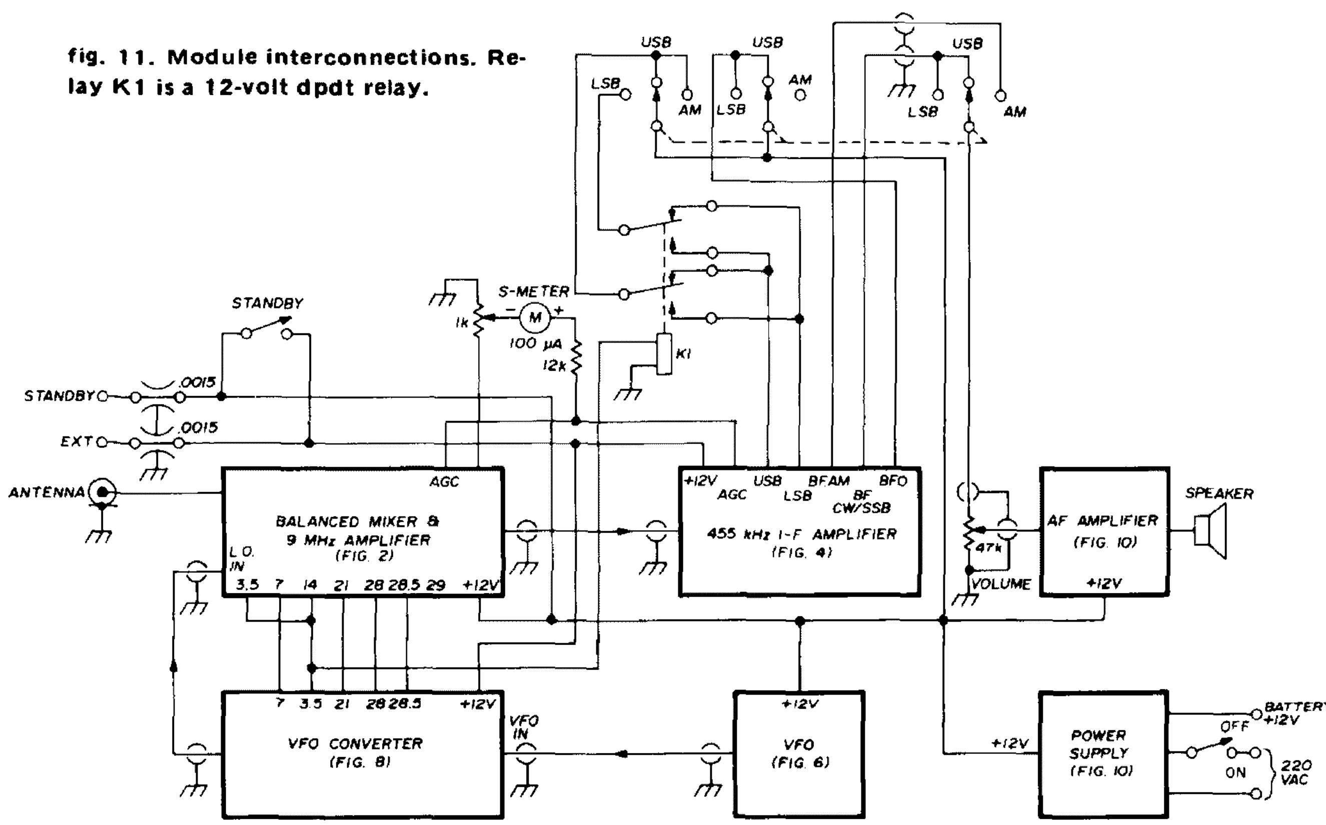


fig. 11. Module interconnections. Relay K1 is a 12-volt dpdt relay.

a 47-ohm, ½ watt resistor across the output jack. With the rf probe at this point and the bandswitch set on 80/40 I measured about 2 volts. The filters and drain inductances may be touched up so that the output stays at about 2 volts at any setting of the bandswitch. After interconnecting the other modules I made final adjustment following normal receiver alignment procedures.

It might be interesting at this point to describe how I measured intermodulation and blocking. For the intermodulation check I set one signal generator for 0.2 μV on 28.7 MHz and measured its frequency 100 kHz above, and 100 kHz below 28.7 MHz. In each case I advanced the amplitude of this second generator until the output S/N ratio deteriorated; when the amplitude was advanced to 70 mV the S/N ratio decreased to 3 dB.

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ham radio

low-cost
1296-MHz preamplifier

Low-noise, high performance 1296-MHz stripline preamplifier uses new Motorola MRF-901 microwave transistor

H. Paul Shuch, WA6UAM, Microcomm, 14908 Sandy Lane, San Jose, California 95124

A welcome side effect of the FCC's recent allocation of frequencies near 960-MHz to the Land Mobile Radio Service is the introduction by semiconductor manufacturers of low-cost, high quality active devices for the low end of the microwave spectrum. This technological revolution has proved to be a *boon to amateur activity in* the 1296-MHz band by bringing state-of-the-art components within the price range of microwave experimenters for the first time. This article presents construction details and performance data on a pair of preamplifiers for 1296 MHz which use the low-noise (under 2 dB), low-cost (under $10.00) Motorola MRF901 microwave transistor. Future developments will undoubtedly bring us a multitude of transistors offering superior performance and/or lower cost.

Numerous previous articles[1-5] have covered the design and construction of high quality preamplifiers for the 1296-MHz amateur band. These amplifiers used outstanding transistors from a number of different manufacturers. Designed primarily for military and aerospace applications, this generation of microwave transistors is characterized by high reliability over a wide temperature range, hermetic construction featuring a ceramic case with gold-plated leads, and (without exception) a cost far beyond the financial resources of the average experimenter.

The new transistors developed for the 960-MHz Land Mobile Band, with plastic cases and tinned-copper leads, are admittedly less rugged than their military predecessors; storage temperature is not above 150°C, breakdown voltages are lower and derating curves are steeper. These devices are, after all, *intended* for commercial service. Their reliability is, however, wholly adequate for an extended life in intermittent amateur service, and their electrical performance at 1296 MHz compares favorably with that of earlier devices costing an order of magnitude more.

design trends

Early solid-state uhf preamplifiers designed for amateur use achieved input and output impedance matching

through the use of pi networks composed of piston trimmer capacitors and slab inductors. This approach, typified by the designs of Katz[1] and Vilardi,[2,3] assured a proper impedance match regardless of the transistor characteristics because of the pi network's unique ability to "match anything to anything." Unfortunately, the pi network's very versatility made it difficult for the amateur who lacked the proper test equipment to know exactly when his amplifiers were tuned for optimum performance. Additionally, extensive tweaking of the input and output circuits made it possible to inadvertently adjust the amplifier into a condition of instability, with the resulting oscillations ultimately destroying the fragile transistor.

The microstripline designs of Donecker[4] and others changed all that. All matching transformers and reactances were etched onto a printed-circuit board, with no tuning adjustments whatever, so there was no need to optimize an amplifier on expensive test equipment. This approach made it practically impossible to destroy a transistor by inadvertently mismatching it. On the other hand, the lack of tuning adjustments made it impossible to compensate for minute differences between components or printed-circuit boards. And, since a board was computer designed to the parameters of a particular transistor, the experimenter had little opportunity to modify the design so he could use a different, more readily available device. In fact, some attempts to substitute transistors on the same circuit board led to a net degradation of the system noise figure to that of a simple diode mixer.

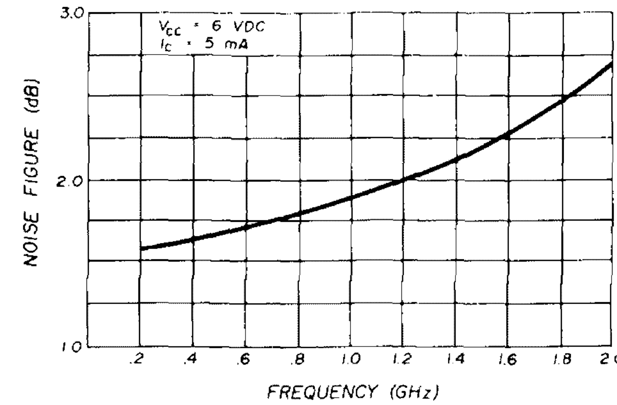


fig. 1. Noise figure vs operating frequency for the Motorola MRF-901 transistor.

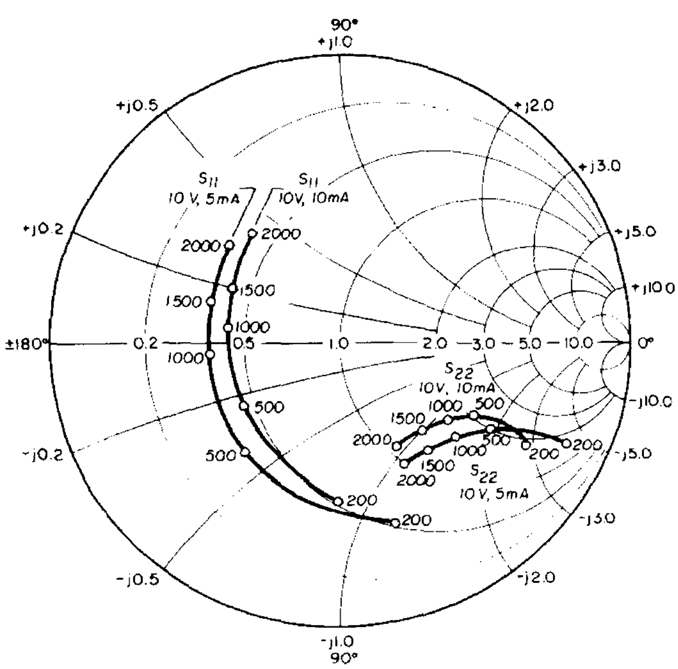


fig. 2. Input reflection coefficient, S_{11} and output reflection coefficient, S22, vs frequency for the Motorola MRF-901.

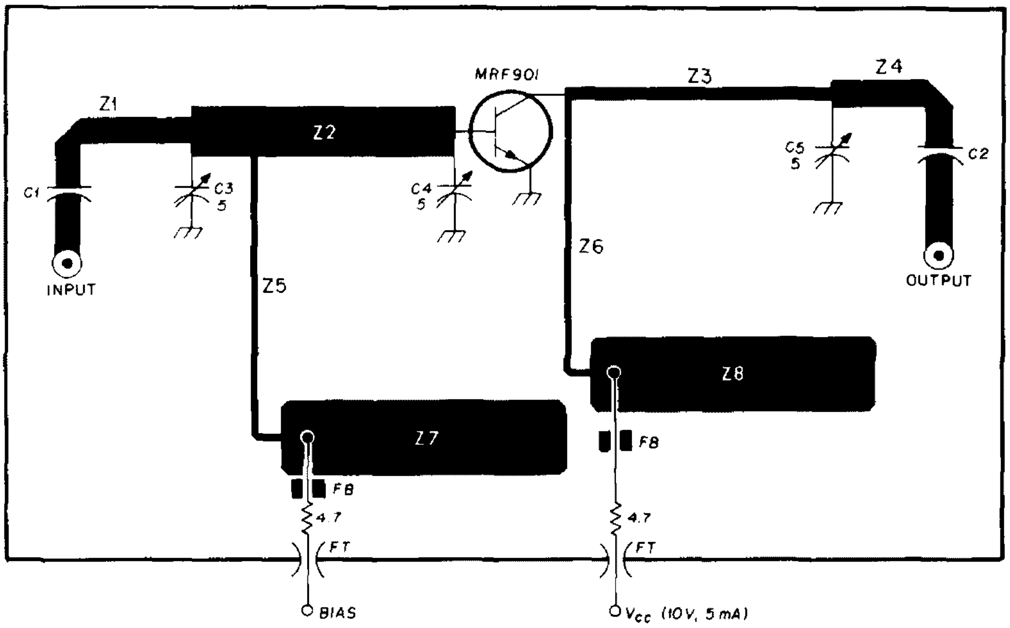
I attempted to rectify these limitations in my preamplifier designs by incorporating "tweaking" capacitors into a printed microstripline design.[5] By allowing one input and one output adjustment per stage, an amplifier is easily optimized, while lessening both the test equipment requirements and the likelihood of circuit instability. The same design approach is used in the preamplifiers presented here.

There is, however, one situation where the pi network matching technique still excels. For optimum system performance, it is necessary to match the input to the *first* preamplifier stage for optimum noise figure, *not* optimum gain. Optimum noise figure is achieved by deliberately and precisely mismatching the applicable input. Most rf transistor manufacturers publish curves or tables of complex impedances for pro-

per input and output power match. Unfortunately, proper *mismatch* information for optimizing noise figure is seldom provided. Thus, it may be desirable to use a wide-range pi network at the input of the first preamplifier stage so that the source reflection coefficient at which noise figure is minimum can be empirically achieved.

transistor characteristics

Fig. 1 shows the manufacturer's claimed noise figure as a function of operating frequency for the MRF-901 transistor. At 1296 MHz the device is capable of delivering a noise figure below 2.1 dB. I have built several noise-figure optimized amplifiers around this transistor, and performance has consistently measured 2.8 dB (with the preamplifier looking into a high-quality double-balanced mixer). The 0.7 dB discrepancy is due, of course, to the familiar second-stage noise contribution, which was summarized quite well in the appendices to Reisert's recent article on 432 MHz preamplifiers.[6] Working through the textbook formula, it is found that the noise-figure optimized MRF-901 stage's intrinsic noise figure is on the order of 2.3 dB. Thus, the losses and mismatch errors inherent in this design still yield performance within 0.2 dB of optimum.



C1,C2	Dc block. 50-150 pF chip capacitor (ATC 100B or equivalent)
C3,C4 C5	1-5 pF piston trimmer (Johannson JMC4642 or Triko 201-01M)
FB	Ferrite bead, slipped over resistor lead
FT	470-1000 pF feedthrough capacitor
Z1, Z4	50 ohm microstripline, 0.1" (2.5mm) wide, any convenient length
Z2	32.4 ohm, quarter-wavelength, microstripline, 0.20" (5mm) wide, 1.15" (29mm) long
Z3	74.8 ohm, quarter-wavelength microstripline, 0.045" (1mm) wide, 1.24" (31.5mm) long
Z5, Z6	Rf choke. 100 ohm, quarter-wavelength microstripline, 0.02" (0.5mm) wide, 1.25" (32mm) long
Z7, Z8	Rf short, 25 ohm, quarter-wavelength open-circuited microstripline, 0.30" (7.5mm) wide, 1.14" (29mm) long

fig. 3. Schematic of the 1296-MHz preamplifier. Capacitor C3 is used only with the noise-matched stage (see text). Quiescent collector current is 5 mA for the noise-matched stage, 10-mA for the power-gain stage. Full-sized printed-circuit layout is shown in fig. 5.

The specification sheets for the MRF-901 list input and output reflection coefficients (S_{11} and S_{22}) in polar form, tabulated for numerous combinations of frequency, power supply voltage and quiescent collector current. The two static conditions of greatest interest to the amateur are 10 volts at 5 mA (for a noise-matched first amplifier stage), and 10 volts at 10 mA (for a gain-optimized second preamplifier stage), as discussed in a previous article.[5] Reflection coefficients corresponding to these bias conditions were plotted on a Smith chart, then connected with a smooth curve as shown in **fig. 2**. The design parameters for the 1296 MHz preamplifiers were determined from interpolation of this Smith chart data.

design procedure

Two preamplifier stages are discussed here, one optimized for noise figure, the other for power gain. The gain-matched second stage was designed along lines analogous to that described in my previous articles. No noise-matched reflection coefficients were available from Motorola for the MRF-901, so a pi network was included in the input to the noise-figure matched stage, as discussed previously.

Fig. 3 is a functional schematic of the 1296-MHz preamplifier stages (the only differences between the stages optimized for noise figure and gain are the quiescent collector current [5 and 10 mA, respectively], and the use of an additional input matching capacitor, C3, in the noise-matched stage).

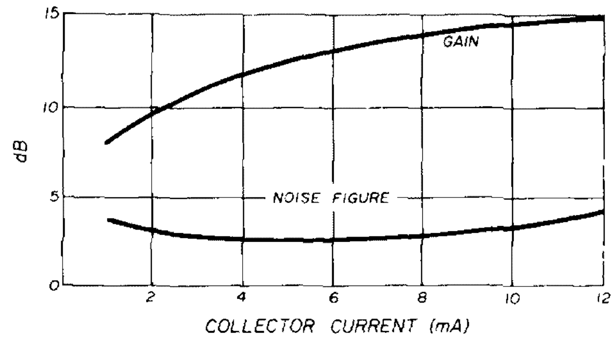


fig. 4. Typical power gain and noise figure of the 1296-MHz preamplifier as a function of collector current.

The system noise figures of several gain-matched 1296-MHz preamplifiers I've built with MRF-901 transistors all measured between 3.2 and 3.5 dB, while yielding 12 to 14 dB of power gain. Thus, for all but the most critical applications, the use of a noise-figure matched first preamplifier stage may not be necessary (see **fig. 4**).

construction

Fig. 5 is a full-sized printed-circuit layout for either preamplifier stage. The only visible difference between the gain-

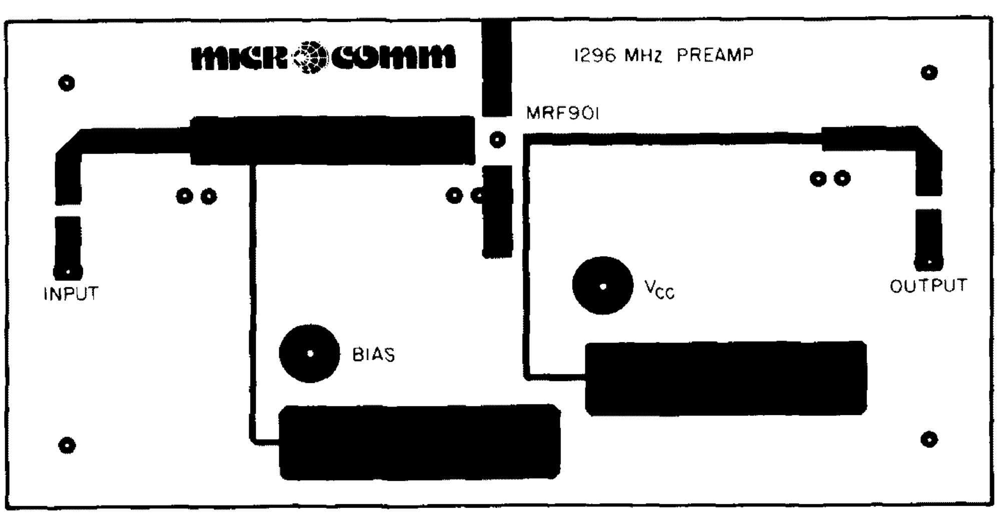


fig. 5. Full-sized printed-circuit board for the 1296-MHz preamplifiers using MRF-901 transistors. Component layout is shown in fig. 3.

matched and noise-matched stages is the incorporation of a variable capacitor, C3, in the noise-matched stage. Note, however, that the quiescent collector current differs between the two stages. More on this later.

The amplifiers are built on 1/16 inch (1.5mm) G-10 fiberglass-epoxy circuit board, double-clad with 1 ounce copper.

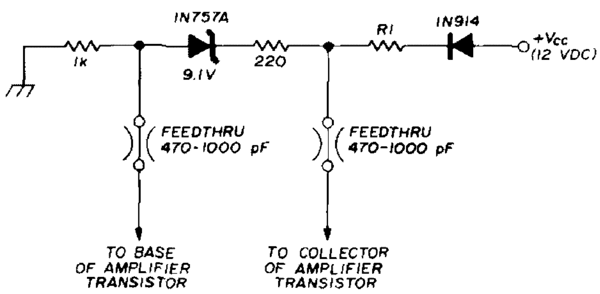


fig. 6. Zener bias circuit, developed by W1JAA, for use with transistor preamplifier stages. For 10 mA collector current, R1 = 120 ohms; for 5 mA collector current, R1 = 240 ohms.

I still receive occasional letters from readers who question my repeated use of glass-epoxy material at 1296 MHz. I will concede that the 0.2 dB excess noise figure above optimum which I have mentioned previously *may* be due to losses in the substrate. However, I feel that 0.2 dB is a small price to pay for the convenience and ready availability of this low-cost printed-circuit material.

Construction of these amplifiers is substantially the same as that of my previously published circuits. If you are unfamiliar with the fabrication of microstripline amplifiers, you are urged to refer to reference 5 for specific construction hints, as well as suggestions for tuneup and operation. That material is not repeated here.

bias circuit

The zener bias circuit introduced by Reisert[6] is not only simpler but also electrically superior to the active bias scheme I used in some of my earlier designs. Reisert's circuit is presented in fig. 6 along with component values for the two required bias conditions. I refer the interested reader to his very fine article for a complete description of the operation of this bias circuit.

One appealing feature of this biasing scheme is that the power supply voltage can be varied upward or downward, as required, to optimize the stage for gain or noise figure. In this manner the performance of the preamplifier can be readily tailored to the requirements of the system in which it is installed.

conclusion

As the commercial microwave communications industry expands, sophisticated amplifiers have come within the reach of the average amateur experimenter. For those who prefer not to build their own equipment, low-cost 1296-MHz preamplifier modules are now available. A commercial version of the gain-matched stage, featuring a guaranteed maximum noise figure of 3.2 dB is currently available for under $40 from Microcomm.*

*For full specifications send a self-addressed, stamped envelope to Microcomm, 14908 Sandy Lane, San Jose, California 95124.

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ham radio

low noise figure 28-30 MHz preamplifier for satellite reception

Inexpensive dual-gate mosfet circuit provides 20 dB gain and typical 1.0 dB noise figure

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Most modern communications receivers for the 28-30 MHz range are designed with dynamic range as the primary goal. Low noise figures generally require more gain ahead of the mixer and hence decrease the dynamic range. Furthermore, noise figures below 10 dB are seldom usable at these frequencies because the local ambient noise usually is the limiting factor for weak-signal detection.

Satellite reception and vhf/uhf converters usually require lower noise figures than the typical communications receiver will provide. Therefore, for these applications it's necessary to install a low-noise preamplifier ahead of the receiver. A suitable preamp which is inexpensive and easy to build is described in this article. It has proven to be a workhorse for many Oscar 6 and 7 operators and has also been used as a low-noise i-f preamp for use with vhf and uhf converters.

Many of the requirements for low-noise preamplifier design were discussed in an earlier article[1] and will not be reiterated here. Field-effect transistors were the prime candidates for this application because they are inexpensive, readily available, and yield low intermodulation distortion.

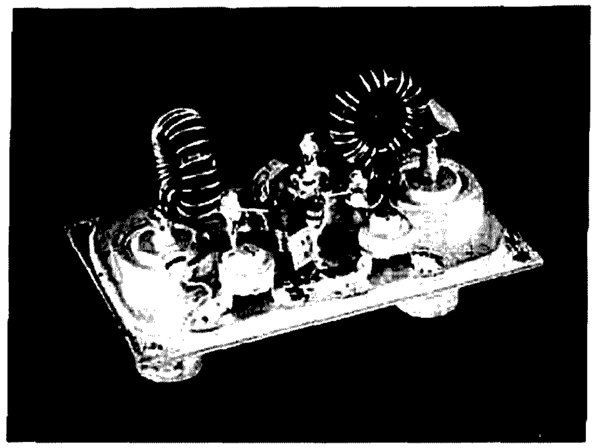
circuit discussion

The three preamplifier configurations shown in **fig. 1** were evaluated to determine which was the most suitable for low-noise operation at 28-30 MHz. Models of each were built and tested. The grounded-source circuit (**fig. 1A**) was discarded since it required neutralization for unconditional stability. Loading the drain circuit helped, but gain had to be reduced considerably before complete stability was obtained.

Next, the grounded-gate circuit (**fig. 1B**) was tried. Although it was stable, gain was lower than expected and the noise figure was slightly higher than the grounded-source configuration. The cascode circuit (**fig. 1C**) seemed to fill the bill; low noise, high gain and stability

were easily obtained using two single jfets *or* a dual-gate mosfet.

Noise figures of 1.5 to 2.0 dB were relatively easy to obtain with this circuit. The optimum source (input) impedance is 2 to 4000 ohms. After calculating the unloaded and loaded Q of such circuits (fig. 2A), I concluded that the noise figure was limited by input circuit losses, not device limitations. I tried the L-network input circuit (fig. 2B) and was pleasantly surprised by the results. The circuit losses were definitely lower. An rf choke which is parallel resonant at the operating frequency was used in the final configuration shown in fig. 2C. This method of matching is one of the secrets to the low noise figure of this preamplifier.



Construction of the low-noise 28-30 MHz preamplifier. Note that the toroidal input and output transformers are installed at right angles to one another.

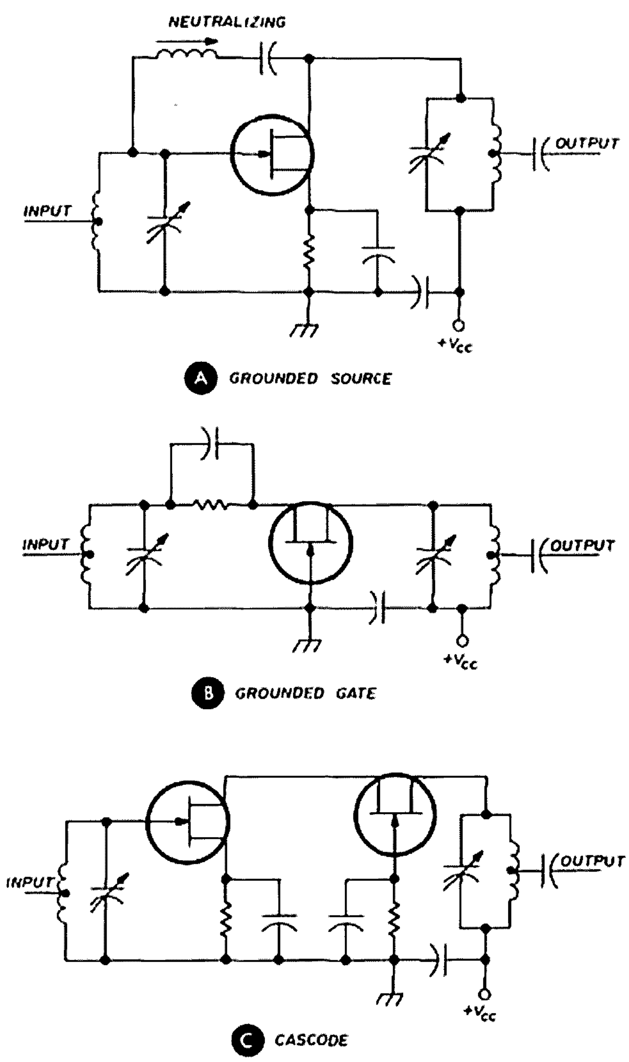


fig. 1. Fet preamplifier circuits which were evaluated for use as a low-noise 28-30 MHz preamplifier. Cascode circuit at (C) provides low noise, high gain and good stability.

device selection

The original circuit (fig. 1A) used a jfet. The Siliconix E300 and popular 2N4416 transistors worked equally well. These devices also worked well in the cascode circuit (fig. 1C). The Fairchild FT0601, a dual-gate mosfet, worked as well in the cascode circuit as the E300 types. Typical noise figures of 1.25 to 1.75 dB were easily obtained but the real surprise came when the inexpensive, diode-protected, dual-gate RCA 40673 mosfet was tried in the circuit. It consistently yielded lower noise figures than any of the other devices I tested: typical measured noise figures were less than 1.0 dB.

circuit description

The original preamplifier circuit had limited publication in March, 1972, and was revised a year later.[2] A complete schematic of the latest version is shown in fig. 3. The input network is as described earlier. RFC1 is parallel resonant at 28-30 MHz; values of 15 and 30 microhenries should be usable with little degradation. Inductors L1 and L2 are low-loss toroidal coils wound on Amidon cores, are small, and provide high

Q. Hence there is very little mutual coupling between the input and output circuits. Capacitor C1 allows optimization of noise figure and usually can simply be peaked for maximum gain. Capacitor C2 tunes the output circuit to the desired operating frequency. The circuit values shown in the schematic will permit operation from about 25 to 35 MHz without any component changes.

Attention is called to the use of a ferrite bead on the gate 2 lead to the mosfet, Q1. Careful testing revealed that most dual-gate mosfets tend to oscillate at uhf, typically at 750 to 900 MHz. This can be readily observed on a uhf spectrum analyzer. The reason for this is easily understood when you consider that there is a small, finite inductance present between gate 2 of the semiconductor dice and the outside of the transistor package. All attempts to bypass this lead were unsuccessful. The ferrite bead, however, did the trick since it essentially puts a lossy element on the lead and suppresses the uhf oscillation.

This same oscillation effect has also been observed on other dual-gate mosfet preamplifiers operating at frequencies as high as 500 MHz. In all cases a ferrite bead on the gate lead solved the problem.

Resistor R2 can be added to the circuit, if desired, to increase bandwidth with a small sacrifice in overall gain. Diode CR1 is an *idiot* diode which prevents damage to the mosfet if the power supply polarity is accidently reversed.

construction

Construction of the 28-30 MHz low-noise preamp is very straight forward and is similar to that used for the low-noise uhf preamp described in reference 1. A small section of double-sided printed-circuit board is attached to the cover of a small aluminum box (such as the Pomona 2417). BNC connectors are used for the input and output. The use of toroidal inductors eliminates the need for any shield if the input and output inductors are located at opposite ends of the box and are oriented at right angles to one another. Further construction details are shown in the photograph.

Although the RCA 40673 is a diode-protected mosfet, care should be exercised when handling the device to prevent any possible damage from static electricity. The best procedure is to first grasp the device by the case and then

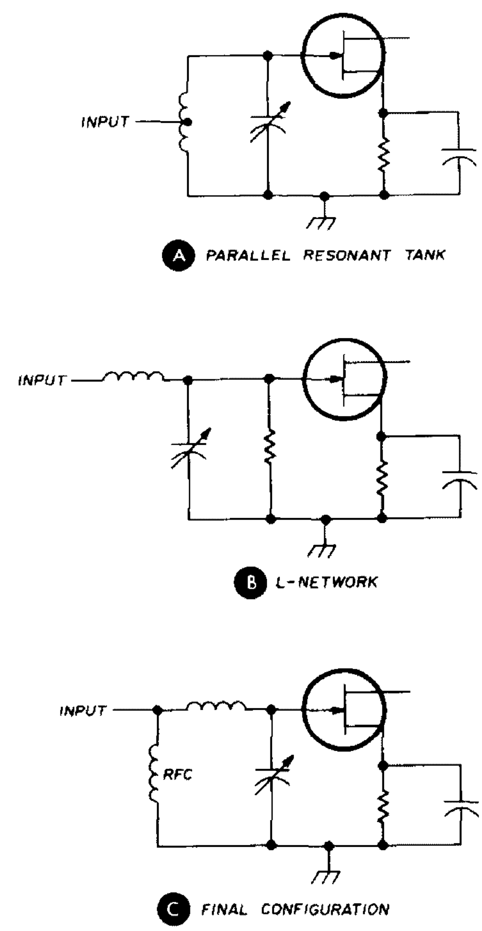


fig. 2. Input circuit configurations. Losses in the tapped, parallel-resonant circuit at (A) measureably increased the noise figure. L-network at (B) resulted in lower circuit loss. The circuit at (C), which uses a parallel-resonant rf choke, provided best performance and was used in the final preamplifier design shown in fig. 3.

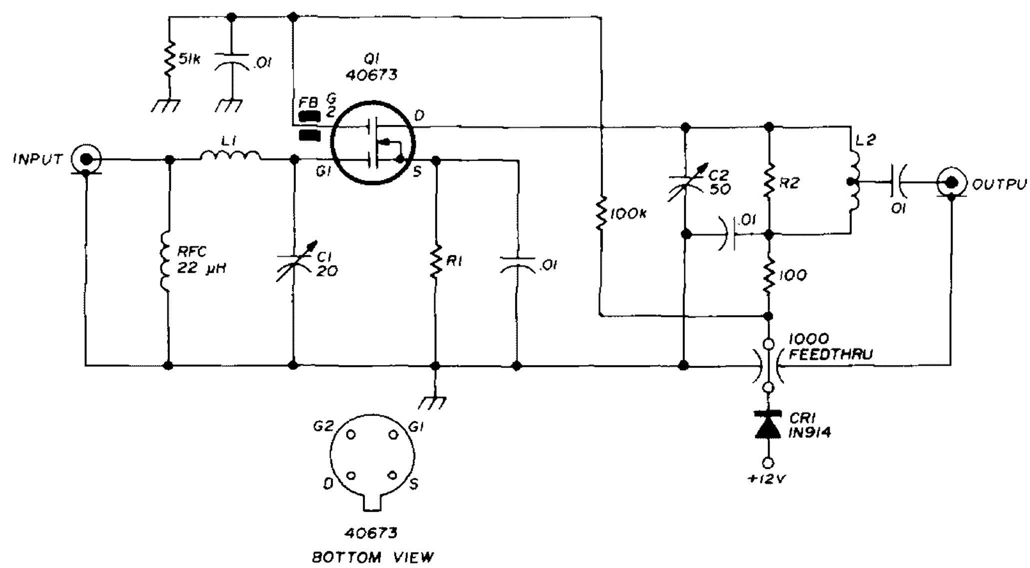
pick up the circuit board. With this technique the potentials are equal and the mosfet leads can be safely soldered into the circuit.

Any low-loss trimmer capacitors may be substituted for C1 and C2. The miniature JFD units I chose were based on

the small size needed to package the circuit in the Pomona 2417 enclosure.

operation and test

For final checkout the input to the preamp should be terminated with a 50-ohm resistor, a noise generator or a 28-30 MHz antenna. The output can be connected to a receiver or i-f amplifier, This preamplifier has been widely duplicated, both for reception of Oscar 6 and 7, and as a low-noise i-f preamp between vhf/uhf converters and a communications receiver. Gain is typically 20 to 25 dB and the highest noise figure I've seen is 1.25 dB. Most preamps have a maximum noise figure of 1.0 dB and some have measured as low as 0.7 dB.



C1	20 pF trimmer (JFD DVJ300 or equivalent ceramic trimmer)
C2	50 pF trimmer (JFD DVJ305 or equivalent ceramic trimmer)
FB	Ferrite bead (56-590/65/3B or equivalent)
L1	25 turns no. 24 (0.5mm) on Amidon T50-10 toroid core
L2	22 turns no. 24 (0.5mm) on Amidon T50-10 toroid core, tapped 7 turns from cold end
R1	150 ohms typical (see text)
R2	2000 ohms typical (see text)

fig. 3. Low-noise 28-30 MHz preamplifier which provides 20 to 25 dB gain and typical noise figure of 1.0 dB uses a dual-gate mosfet in the cascode circuit.

The current drawn from the 12-volt power supply should be checked for proper operation: 3.0 to 7.0 mA is fine. If the current is too low or too high the value of R1 should be raised or lowered accordingly.

The only adjustments are to peak the input and output capacitors (C1 and C2, respectively) for maximum output. If a noise generator is available, C1 can be adjusted for lowest noise figure. However, as noted previously, in most cases good low-noise performance can be obtained simply by adjusting C1 for maximum gain.

The extremely low cost of this unit makes it a real bargain. Once you have one in operation, you'll wonder how you ever did without it.

Special thanks go to all those who have duplicated this preamp and encouraged me to write this article. I hope its use will improve your station's performance.

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ham radio

bfo multiplexer

for a multimode detector

Cmos logic oscillator deck and phase-locked loop are combined for a multimode detection system

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For several years I've been experimenting with solid-state circuits to replace various receiver subsystems at my station. Most of the stages of my venerable HRO-50T have been rejuvenated in this fashion. Installation of this multimode detection system was a major step toward providing the performance and operating conveniences demanded of today's best receivers.

In the bfo multiplexer a single IC functions as an upper-sideband crystal oscillator, lower-sideband crystal oscillator, tunable bfo for CW, or as a limiter of the i-f signal for fm or synchronous a-m reception. The desired oscillator (or the limiter) is gated on by grounding its digital contol line. Multimode reception results when the multiplexed output of the oscillators and limiter is applied to a product detector. Fm signals can also be demodulated if the NE561 is used as a combination product detector and phase-locked loop discriminator.

Construction details are provided for a 455-kHz multimode detection system using the NE561. The IC module is packaged inside the original bfo coil's shield can for ease of installation. It is most economical, considering the features it provides, since surplus FT-241A crystals may be used and the ICs are not expensive.

cmos logic oscillators

Crystal-controlled oscillators using linearly biased cmos logic elements are described in RCA's applications literature.[1,2] In the basic circuit, shown in **fig. 1**, R1 biases the logic element as an inverting linear amplifier. The pi-type feedback network provides 180 degrees phase shift at the crystals's parallel resonant frequency to produce oscillation. The crystal sees a load capacitance equal to the series equivalent of C_s and C_t.

With regards to stability, the cmos oscillator circuit shown here is competitive with good discrete component de-

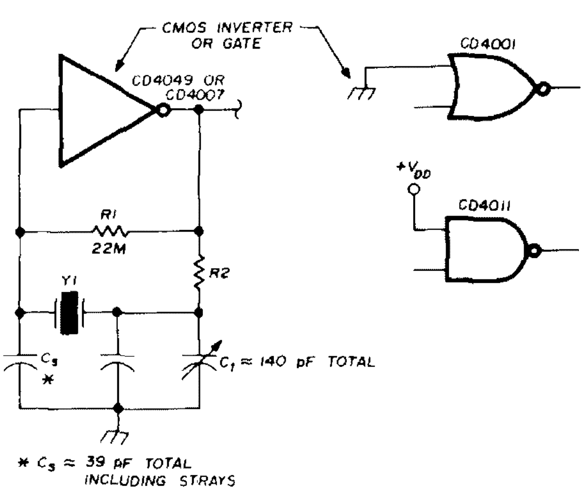


fig. 1. Basic crystal oscillator circuit using a cmos inverter or gate. Crystal Y1 is 10 MHz or less.

signs. The prototype of a 3.3 MHz oscillator that I am using in an industrial application, for example, is built around a high quality crystal and drifted less than 20 Hz between 0°C and 70°C without any temperature compensation. Power supply sensitivity has been reported as 3.5 ppm for a 25 per cent change in supply voltage.[1] Power drain, although proportional to frequency, is extremely low and a wide range of supply voltages can be accommodated.

Resistor R2 in **fig. 1** is the only component whose design value must change with frequency. At high frequencies it must be reduced or eliminated since the attenuation and phase shift it introduces increases with frequency.

multiplexer circuit

The idea which gave birth to the bfo multiplexer was that of obtaining four amplifier-oscillators, each with an on-off control, from a single quad 2-input NOR gate package and combining their outputs in a *resistive summing network*. The resulting circuit is shown in **fig. 2**. The supply voltage, V_{DD} should be between 3 volts and 15 volts. Regulation is necessary only to avoid exceeding the 15-volt upper limit. With a 9-volt power supply and the component values shown, a 770 mV p-p square wave output was observed. If you must use different summing network component values, use about 1000 ohms per volt of the supply in series with each output of the CD4001 IC to avoid overloading it.

Transformer T1 in the tunable oscillator circuit is a Radio Shack transistor oscillator coil. However, any high-Q coil-capacitor combination which resonates at the i-f will do. The stability of this circuit, particularly warm up drift, was noticeably better than that of the original circuit.

The input from the i-f amplifier should be at as high a level as possible for good limiting. This consideration should present no problem since the high input impedance of the CD4001 will not load a high-Q tuned circuit. Sensitivity for full output is about 300 mV p-p. The input level is adequate if a rounded square wave is obtained at the output with an a-m signal tuned in.

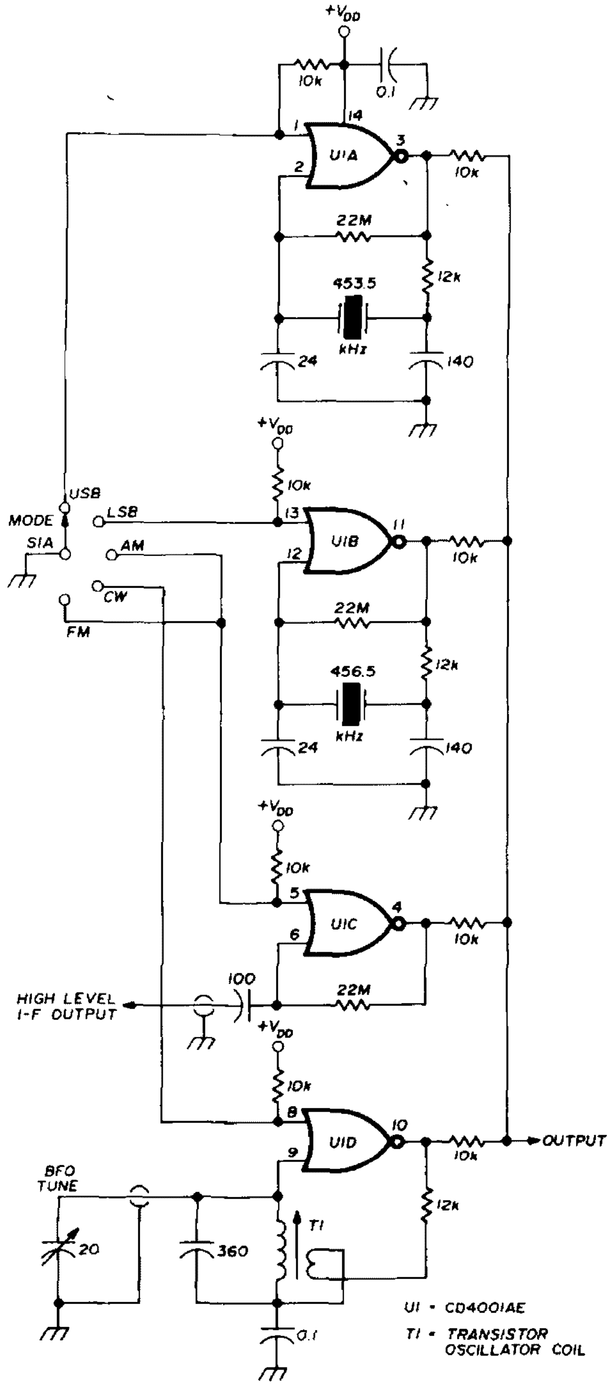


fig. 2. Schematic diagram of the bfo multiplexer. The supply voltage, V_{DD}, should be between +3 and +15 volts.

detector circuit

The bfo multiplexer provides a-m detection capability for a receiver equip-

ped with only a product detector while simplifying mode switching. In this application it replaces the original bfo. I first used the bfo multiplexer with a MC1496G product detector.[3] The results were excellent. Gain and dynamic range of this device exceed that of the NE561. A further advantage was that both the detector and bfo could use the same supply voltage. On the other hand, the gain and dynamic range were not needed and fm detection capability was desired.

When used with the NE561, a 90 degree phase shift network is required in series with the output of the bfo multiplexer as shown in fig. 3. During a-m reception the network compensates for the fact that the NE561 locks up in quadrature with the signal at its phase detector's input.

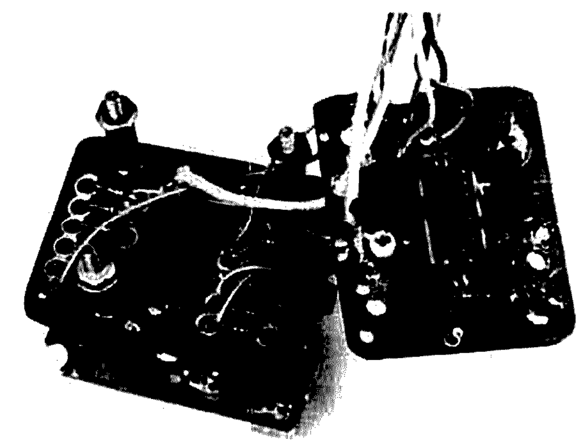
The i-f input level to the NE561 should be held to about 100 mV rms for minimum distortion. If this is done the audio output level will be at least half that for narrow-band fm, about the same for ssb and CW, and at least twice for a-m if both sidebands are passed by the i-f filters. Note that the fm audio output level is proportional to the percentage deviation and cannot be increased by increasing the signal level. The two 0.004 μF capacitors limit the audio bandwidth to about 4 kHz.

The vco output of the NE561 is a 0.6 volt p-p square wave at the a-m carrier or bfo frequency. It appears to be ideal for use as an input to a digital dial adapter for accurate received frequency measurement.

construction

The multimode detection system was built on three 2-1/2 X 1-3/4 inch (57x45mm) copper-clad epoxy boards. The boards were stacked with phenolic posts, then slid inside the original bfo coil's shield can. The top board was bolted to the top of the shield can. Since the boards fit snugly inside the can, the complete assembly meets mechanical stability requirements. There is still room in the shield can for a fourth board containing a cable connector, if you are so inclined.

The crystals and associated components are mounted in an octal socket on the top board. The oscillator coil and its associated components are mounted nearby. The top of the can is cut out to allow access to the crystals and the coil's adjustment screw without disassembly. Space could have been found for two trimmer capacitors for the crystal oscillators, had this been considered necessary.



Closeup view of the detector board, right, and the first multiplexer.

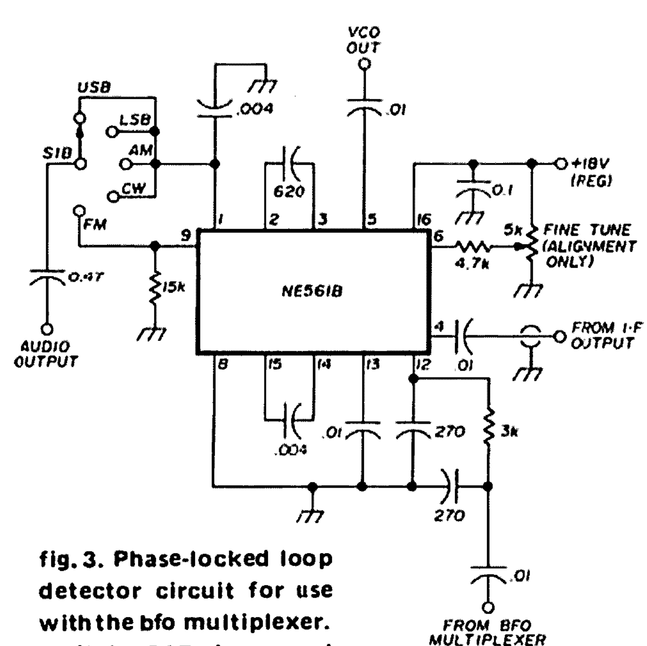
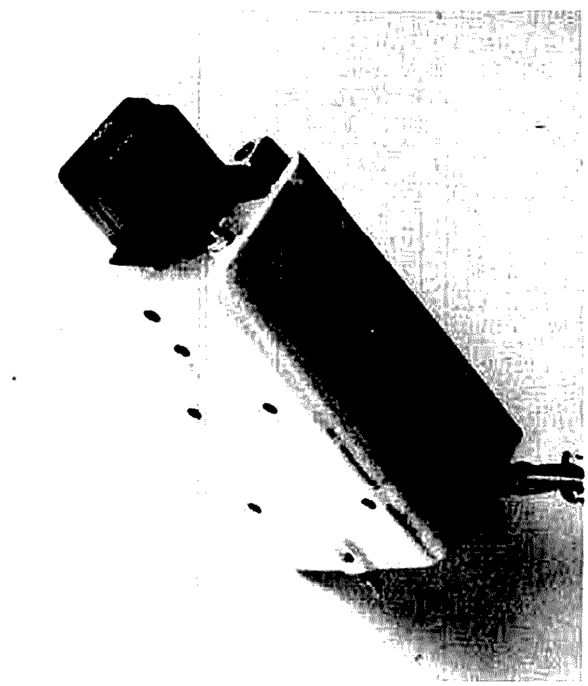


fig. 3. Phase-locked loop detector circuit for use with the bfo multiplexer. Switch S1B is ganged with the bfo multiplexer mode switch, S1A, in fig. 2.

The CD4001AE is mounted in a 14-pin DIP socket on the middle board. The two rows of resistors which flank the socket in the photograph are the pull-up resistors, summing resistors, and 22M feedback resistors. The pads were made with a Vector pad cutting tool.

The layout of the detector board can also be seen in the photograph. The fine-tuning control is adjacent to pin 16 of the IC. With the bfo multiplexer in-



The completed multimode detector is housed in the original bfo shield can of the author's receiver.

stalled in the receiver, it can be reached through a hole drilled in the chassis. Alignment consists of setting the free-running frequency of the NE561 (bfo multiplexer off or disconnected) to within 200 Hz of 455 kHz using this control.

operation

This detector system may be operated just like any other product detector-bfo combination. Most of the time you will be completely oblivious to the fact that there is a phase-locked oscillator in the system. However, when the limited i-f signal is applied to the NE561, several peculiarities may become evident. If conditions are just right, a coded CW signal may be demodulated in the a-m position as a series of chirps as the vco falls into and out of lock on each dit or dah. Another anomoly results when an interfering heterodyne is coincident with one of the voice sidebands of an a-m signal. Depending on their relative strengths, the loop may lock up on the interference, rather than the a-m carrier.

Excessive lock range can also be a problem. If strong signals hold the vco until you have tuned well past them, seemingly putting hysteresis in the tuning mechanism, a circuit modification may be in order. I found that a 22k resistor between pin 7 of the NE561 and its power supply reduced the lock range to ±6 kHz. A less drastic solution is to switch to the CW oscillator those few times it is necessary to bring the vco back on frequency.

These phenomenon are simply quirks and are not reasons for not taking advantage of phase-locked loops for multimode detection applications. Other types of detectors have other problems under the same conditions.

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ham radio

high-performance balanced mixer for 2304 MHz

Construction details for a hot-carrier diode mixer using stripline design that can provide a 7-dB noise figure at 2304 MHz

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In a previous article about a 1296 MHz balanced mixer[1] I mentioned using a similar circuit for 2304 MHz. This elicited several requests for information about the 2304 MHz mixer shown in **fig. 1**. However, since I felt that the ceramic board construction was beyond the facilities available to most amateurs, I was unable to respond. Since I was curious to see if simple construction techniques would still work at 2304 MHz, I finally made the version shown in **fig. 2** which uses an ordinary printed-circuit board. To my surprise, the second version exhibits *better* performance than the "professional" one.

The basic design, shown schematically in **fig. 3**, is identical to the 1296 MHz mixer described in *QST* and consists of a 3 dB quadrature-hybrid coupler, quarter-wave stubs for bypassing, and a low-Q pi network for i-f impedance matching. The major difference is the use of 1/32 inch (0.8mm) double-clad, G10 epoxy-fiberglass circuit board, rather than 1/16 inch (1.5mm), to maintain a reasonable aspect ratio. The line dimensions shown in **fig. 4**, however, are more critical because of the thinner board and higher frequency.

I cut the printed-circuit mask directly to size on Rubylith* with a knife and a ruler so it should be possible to

duplicate the layout with tape or by cutting the pattern directly into the copper and peeling away the excess. Dimensions A through E, and especially line widths A and B, should be within 0.005 inch (0.1mm) of the values shown for best results.

construction

Construction is exceedingly simple, requiring only a drill and a vise. The circuit board is attached to a shelf in a 2-3/4 x 2-1/8 x 1-5/8 inch (70x54x41mm) Minibox. The shelf dimensions are shown in fig. 5. Use enough screws to keep the board flat against the shelf and to provide a ground path for capacitor C4. Metal screws have no effect if they are kept away from the microstrip transmission lines. Approximate placement of the screws can be seen in fig. 2.

Critical parts are the blocking capacitors, C1 and C2, the connectors, and, of course, the mixer diodes. The blocking capacitors should be low-loss chip capacitors, preferably physically small to maintain a low vswr. The same considerations, loss and vswr, also apply to the connectors; SMA type connectors similar to the ones shown are available very reasonably from E. F. Johnson.

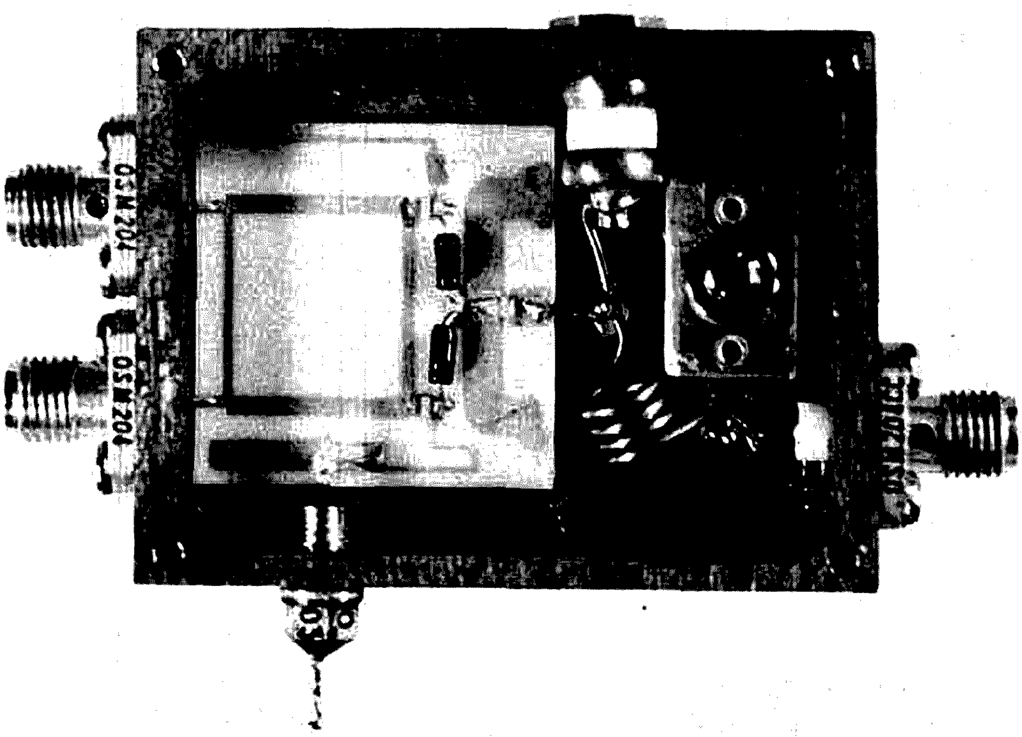


fig. 1. Original balanced mixer for 2304 MHz used ceramic substrate. Small blobs are silver paint used for fine tuning.

adjustment

Adjustment is the height of simplicity. A two-meter converter is connected to the i-f output, and a *clean* 1 to 2 milliwatts at 2160 MHz is fed into the LO connector. Apply about 1.5 mA of bias current to the diodes and apply a

*"Rubylith" is a trademark of Ulano Co.

moderately strong signal at 2304 MHz. When the bias current and i-f trimmer capacitor are adjusted for maximum signal, the tune-up is completed.

performance

A testimonial to the performance of this mixer is that, at the Eastern VHF/UHF Conference earlier this year, the original version won a certificate for the lowest noise-figure 2304 MHz converter. The measured noise figure was 7.1 dB (i-f noise figure, 2 dB), but, in the interest of fairness, W1JAA insisted that 3 dB be added because of the poor image rejection, so the final noise figure was 10.1 dB. The new version described in this article measures *2 dB better*! The mixer has a conversion loss of 6.1 dB.

Noise figure, both theoretical and measured, is the sum of conversion loss plus i-f noise figure. With a 1 dB noise figure at two meters, an overall noise figure of approximately 7 dB could be realized. This should compare favorably with most other mixers at this frequency; W2CQH's interdigital mixer[2] may be better, but it is also more complex. Isolation between the local-oscillator to the rf ports measures 22 dB; this indicates that the hybrid coupler is working properly and that the diodes are well matched to 50 ohms (any power reflected from the diodes will decrease the isolation).

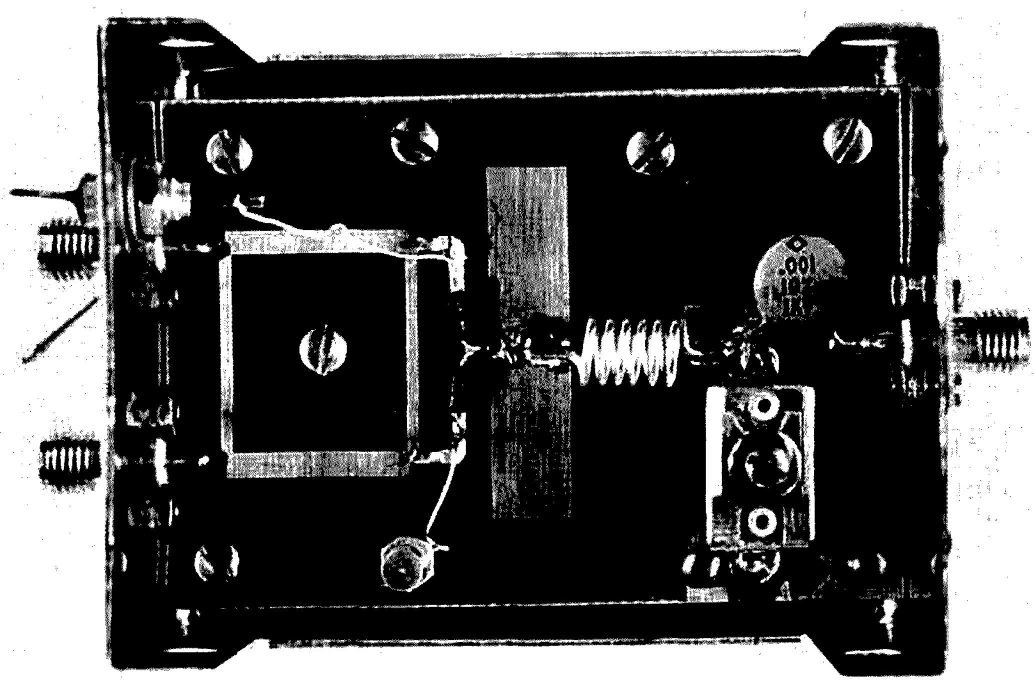
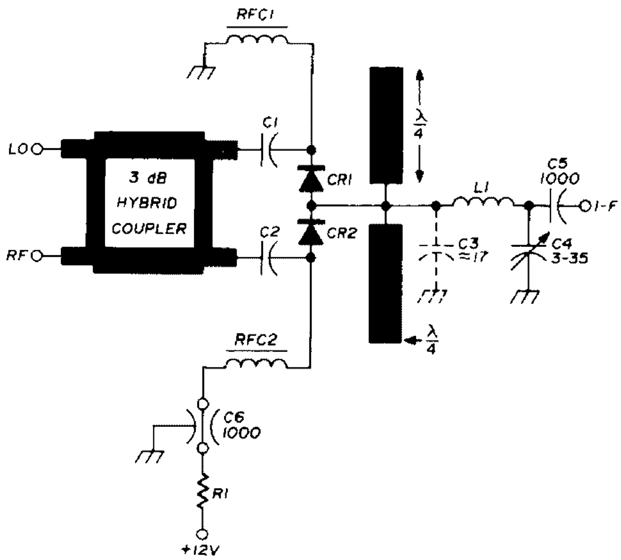


fig. 2. Balanced mixer for 2304 MHz using G10 printed-circuit board. No tuning is necessary.

dc bias

A few words about the value of dc bias in mixers may be in order. I have not seen any other amateur designs which include biasing of the mixer diodes, whereas it is fairly common practice in commercial mixers. Prior to building my first balanced mixer (for 1296 MHz), I made some measurements of diode impedance versus rf signal power (LO) and bias current. The conclusion was that, with dc bias, less



C1,C2	50 to 1000 pF chip capacitor (not critical, see text)
C3	stray capacitance of quarter-wavelength stubs (≈17 pF)
C4	3-35 pF mica trimmer (Arco 403)
C5	1000 pF disc ceramic
C6	1000 pF feedthrough bypass
CR1, CR2	hot-carrier diodes rated for 3 GHz operation (H-P 5082-2535, H-P 5082-2565, Alpha D5501, matched pair not essential)
L1	6 turns no. 18 (1.0mm), 3/8" (9.5mm) long, wound on 3/16" (5mm) mandrel
R1	5k to 15k, vary to set diode bias current
RFC1,2	1 turn no. 30 (0.25mm), wound on 1/16" (1.5mm) mandrel
J1-J3	SMA connector (OSM 215 or E. F. Johnson 142-0297-001)

fig. 3. Schematic diagram of the balanced mixer for 2304 MHz. Dc bias is provided to the diodes to minimize LO requirement. Full-size printed-circuit layout is shown in fig. 4.

LO power is required to raise the diode impedance to 50 ohms, and the impedance is less sensitive to the drive level.

We are, in essence, substituting readily available dc power for hard-to-get rf power! This is borne out by the fact that my mixers work fine at LO levels of one milliwatt, while other, similar designs[3,4] specify around three milliwatts. Also, Tilton recently mentioned[5] several balanced mixers which were rejected by *QST* – their poor performance was apparently due to lack of LO power. Addition of dc bias might have helped significantly.

Bias current is not critical (in this mixer minimum conversion loss was achieved with 1.8 mA of diode current), but it can be varied from 1.0 to 2.6 mA with only a 1 dB increase in loss. With no bias, however, conversion loss increased to 13 dB. Small changes in LO power, simulating normal drift, also had a minimal effect.

One final advantage of dc biasing may be to force the diodes to operate at the same current, and hence at similar impedance, for no attempt at diode matching or individual tuning was made in order to achieve the stated performance.

balanced-mixer design

I have received several queries about designs for other frequencies. These can be made by direct scaling from this design or the 1296 MHz version, following these guidelines:

1. Dimension A is for a characteristic impedance, Z_o = 50 ohms, and B for Z_o = 35 ohms. No change is required if the same board material is used.

2. Dimensions C and D are one-quarter wavelength long at a frequency halfway between the signal and LO frequencies (2232 MHz in this case).

3. Dimension F is one-quarter wavelength long at the signal frequency (2304 MHz).

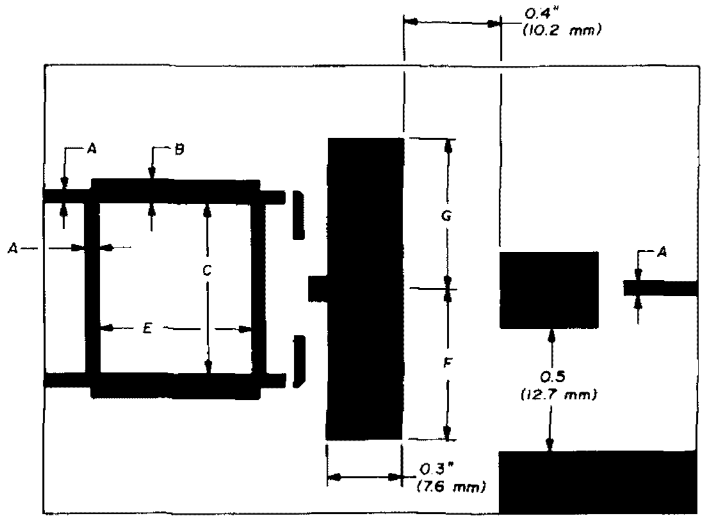
4. Dimension G is one-quarter wavelength long at the local-oscillator frequency (2160 MHz).

5. Wavelength in microstrip line is a function of Z_o, so use a graph of Z_o and λm if you change board material (different dielectric constant).

6. Dimension E is D minus A.

7. Use a low Q i-f circuit which matches the impedance of the diodes (100 to 200 ohms at the i-f).

If the same board material is used (1/32 inch G10) scaling dimensions C, D, F and G directly by the ratio of the frequencies will work fine. As mentioned above, good isolation from the LO to rf ports is an indication that everything is working properly; measure this after all other adjustments have been made.



	dimensions	
A	0.053"	(1.3mm)
B	0.096"	(2.4mm)
C	0.696"	(17.7mm)
D	0.678"	(17.2mm)
E	0.625"	(15.9mm)
F	0.610"	(15.5mm)
G	0.650"	(16.5mm)

fig. 4. Full-size printed-circuit board for the 2304-MHz balanced mixer. Microstrip dimensions shown at right may be scaled to other frequencies (see text).

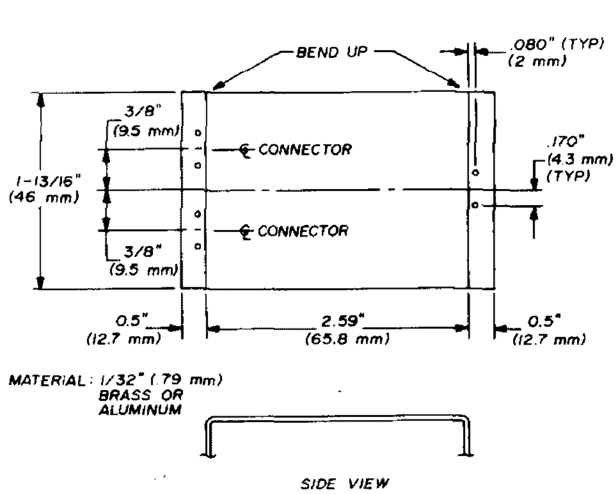


fig. 5. Shelf dimensions for installing the balanced mixer in a standard Bud CU-3000A Minibox.

conclusion

The balanced mixer will provide excellent performance which is easily duplicated and maintained. Use of a low-loss filter at the input is recommended to improve image rejection; there are no other responses if a clean local-oscillator signal is used. A preamplifier may be used for even lower noise figure, and the mixer will provide a low vswr load for the preamp to help stability. However, unless you really need the small (and costly) improvement a preamp provides, the mixer alone has two major advantages — it is much harder to burn out, and it will *not* oscillate.

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ham radio

satellite receivers

for repeaters

Adding satellite receivers to your repeater won't bring in OSCAR 7 — but it might improve your terrestrial coverage

Fred Studenberg, Jr., WA4YAK*

Many users of low-powered handie-talkies have found that the theory of reciprocity proves to limit their ability to communicate through a repeater. Reciprocity implies that, with equal receiver sensitivity and antenna gain on both ends of a path, equal transmitter power is necessary for both stations to hear each other with the same signal strength.

Generally, a repeater's transmitter will have a 17 to 23 dB power advantage over a two-watt portable, effectively reducing the repeater's usable sensitivity by the same amount. This is the typical case of hearing the repeater full quieting but not being able to access it, even though the repeater's receiver is very sensitive.

Assuming that it is desired to retain the present transmitter coverage, there are several ways to balance out the transmit-receive range. In the case of a split-site system, the existing receiver can be relocated at the highest elevation possible. This is, of course, assuming that the repeater was set up initially with equal transmitter and receiver

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heights. Many repeaters were originally designed for use with mobiles that ran 30 to 60 watts and tended to hear as well as they transmitted to the repeater.

satellite receivers

If you operate a single-site system, or have run out of tall buildings and towers, there is another solution to the problem. The concept of satellite receivers has been applied successfully for some years by commercial users and some advanced repeater groups. These remote receivers, located in areas away from the main receiver, relay the weaker incoming signals by vhf or uhf links to the main transmitter.

Ideally, a satellite receiver system should have circuitry that will automatically select the audio from the receiver that is hearing the best quieting signal. However, our repeater group has found that very elaborate methods are not really needed. A typical fm receiver will yield usable audio from any signal that can open the squelch, even though the receiver may only be quieted by 10 or 15 dB. With this in mind, a four-channel scanner receiver was built for the 450-MHz band. **Fig. 1** shows how this is used in our satellite system.

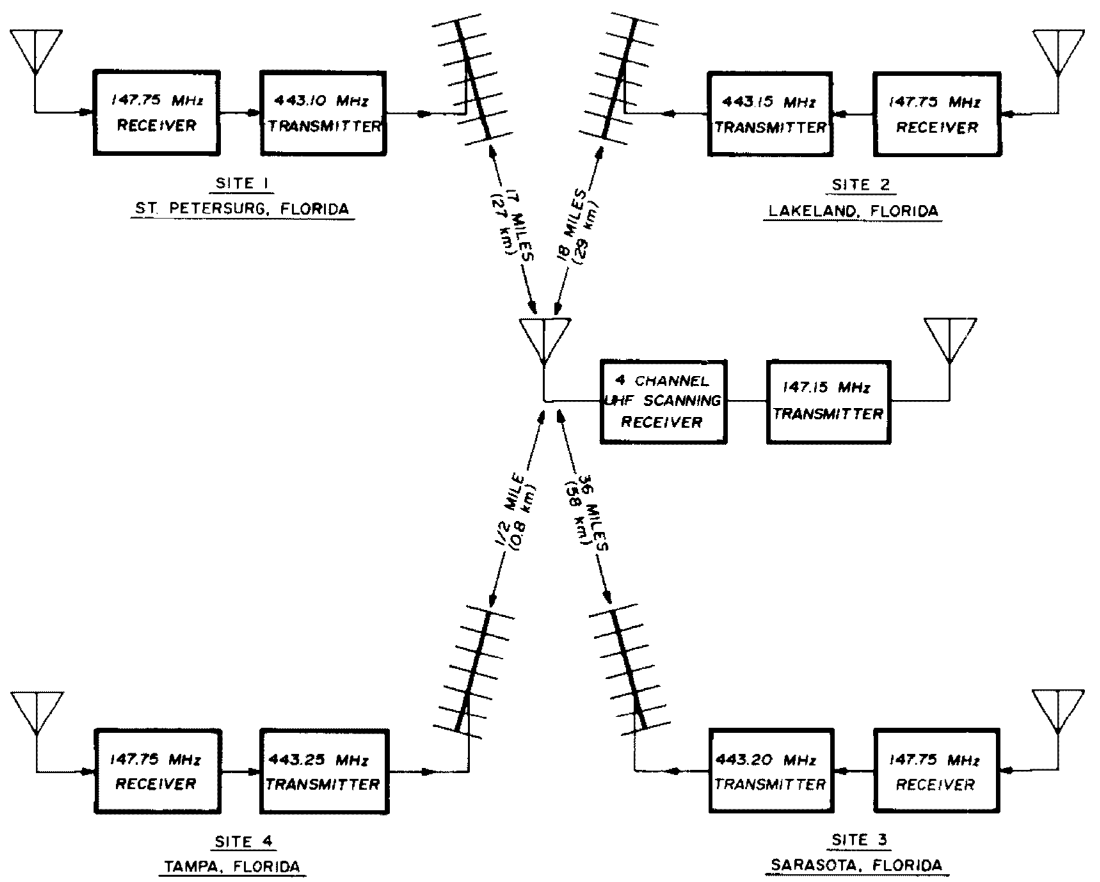

fig. 1. Satellite receivers expand the two-way coverage area of repeater WR4ALT in Tampa, Florida.

The receiver will lock up on the first link transmitter it hears during the scan process. The system tends to seek out the link transmitter that is carrying the best signal, since a link carrying a "chopping" mobile signal will also be "chopping" into the link receiver. The scanner link receiver will tend to unlock and lock up on a solidly keyed link

from a receiver that is hearing a full quieting signal. To avoid wear on the link transmitter's keying relay and also to provide security to the link receiver, the actual presence of a signal is indicated by a 10-kHz subcarrier modulated on the link transmitter. The squelch, and consequently the scanning action, is dependent upon the presence of the 10-kHz subcarrier.

problems encountered when trying to make one-to-one comparisions in strengths of the remote receivers.

To satisfy FCC requirements, each satellite system is licensed as an auxiliary link station, and the link transmitters are identified continuously by a 4-kHz MCW signal. This is filtered out in the link receiver, and the main transmitter is identified by its normal identi-

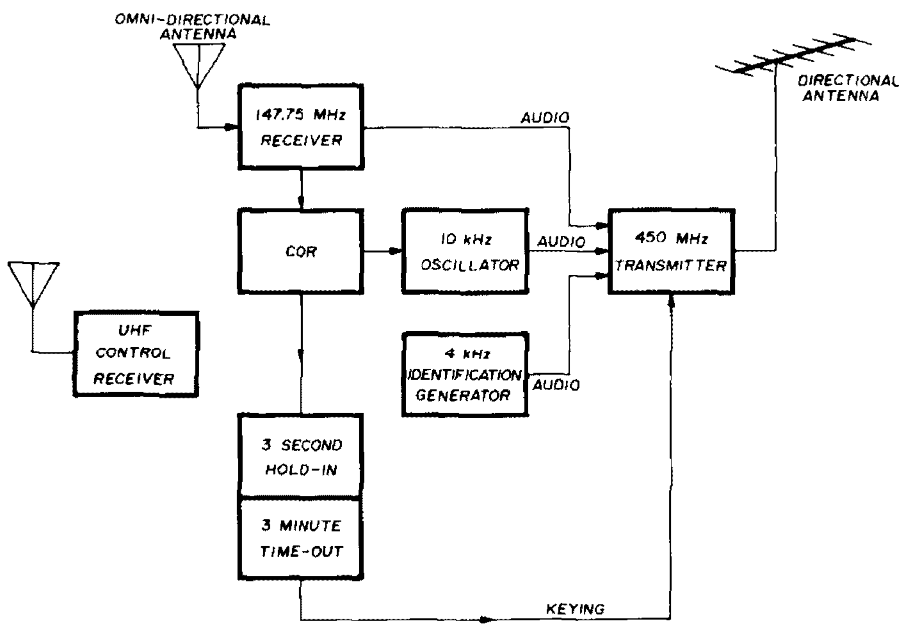

fig. 2. A typical remote receiver installation. The 10-kHz sub-carrier ensures that 450-MHz signals from other sources besides the satellite receivers will not be repeated on the 147-MHz output frequency. It also serves as a "tail-less" squelch since the 450-MHz transmitter is held on by the 3-second hold timer.

Referring to **fig. 2**, the satellite receiver's COR keys the 10-kHz oscillator and also the link transmitter, but through a three-second hold-in timer. This effectively eliminates the squelch tail from the link receiver since it resumes scanning when the subcarrier is removed, even though the carrier is still present. A 10-kHz tone was chosen to allow fast detection time. A Signetics NE567 phase-locked loop tone decoder is used, and it can recognize the subcarrier in less than 1 millisecond. A scan search rate of 25 milliseconds per channel is used, allowing a worse case delay of about 80 milliseconds when selecting different satellite receivers. This has proven to be a very acceptable method of voting without any of the

fication circuits. Regulations require control of the link transmitter. In our case, a common radio control frequency is used for all five sites, simplifying the control operator's equipment requirements. Link frequencies in the 450-MHz band were chosen since monitoring of the link transmitters is not required. Link operation on 220 MHz would require four-frequency monitoring capability.

Selection of the remote sites should be dictated by "dead spots" in the repeater's current coverage area. Naturally, high antenna heights are desirable, but even low antenna satellite receivers will do wonders for filling in receive coverage for the handie-talkie users.

ham radio

crystal discriminator

for vhf fm

High-performance crystal discriminator for vhf fm can be built around an inexpensive third-overtone CB crystal

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Many amateurs, and professional design engineers as well, have attempted to design single-conversion fm receivers only to scrap their brainchild because of low recovered audio and the resulting noisy audio output with poor squelch action. A single-conversion vhf fm receiver must have an intermediate frequency of 5 MHz or higher to provide adequate image rejection. The percentage of deviation is quite small at the higher intermediate frequencies so ICs utilizing quadrature detection may provide less than a millivolt of recovered audio. A simple solution to this problem is to increase the Q of the quadrature coil or use a crystal in its place. Unfortunately, however, these ICs become very unstable and impossible to tame when the Q of the quadrature resonator is increased.

crystal discriminators

Most of the *novel* discriminator circuits, such as the pulse-counting or digital type and transformerless type, as well as the conventional Foster-Sealey, Round-Travis and the ratio detector, do not provide enough recovered audio to be useful at the higher i-f frequencies. Crystal discriminators are quite popular in some of the more sophisticated commercial radios. In fact, Motorola uses a two-crystal discriminator to obtain a plus and minus voltage swing for automatic frequency control in some radios. The crystal discriminator was mentioned briefly in *QST* but no specific values were given.[1]

Most crystal filter manufacturers market a crystal discriminator which

complements their crystal filter line so they are very reluctant to disclose the internal workings of these devices. However, several hours of investigation turned up a specification sheet from C-F Networks, which included all the necessary data to design other crystal discriminators and tune them.

Excellent 9-MHz crystal filters (manufactured by KVG) are available from Spectrum International.* In addition, monolithic crystal filters and monolithic filter elements are manufactured by Piezo Technology, Inc.† Spectrum International also markets an excellent

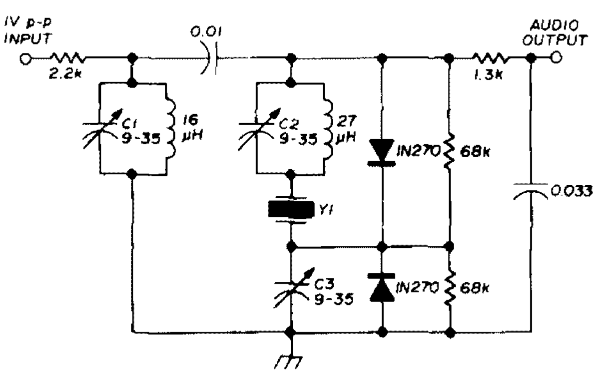

fig. 1. Low cost fm crystal discriminator uses third-overtone channel-1 CB crystal (26.965 MHz) at its 9-MHz fundamental. Response curve of this circuit is plotted in fig. 2.

KVG crystal discriminator for approximately $25.

Quite a few CB crystals are available on the surplus market and in various ham-shack junkboxes so one of these 27 MHz, third-overtone crystals was put into a breadboard circuit and carefully tested at its fundamental frequency, 9 MHz, for modulation acceptance, recovered audio and distortion. Modulation acceptance was sufficient for 5 kHz deviation but more than 7 kHz deviation caused slight audible distortion. A

*Spectrum International, Box 1084, Concord, Massachusetts 01742.

†Piezo Technology Inc., Post Office Box 7877, Orlando, Florida 32804.

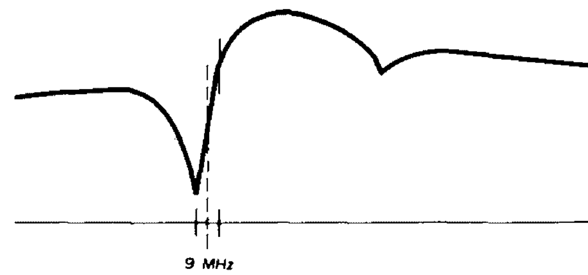

fig. 2. Response of the 9-MHz crystal discriminator. The linear section of the response curve just above the fundamental resonant frequency is the portion actually used. Good filtering is required ahead of this discriminator or serious audio distortion will result.

12 kHz bandpass filter would provide a perfect match for the discriminator.

The crystal used for the tests was a channel-1 transmit crystal for a Heathkit Model CB-1. It is a 26.965 MHz, third-overtone crystal identical to those used in hundreds of other CB radios. This particular crystal is in the larger HC-6/U holder but many are in smaller holders.

the circuit

Fig. 1 is the schematic of the crystal discriminator circuit. Capacitor C3 is adjusted for zero voltage with an unmodulated carrier at center frequency but this setting may not hold for all crystals, and it is possible to obtain good performance and good audio re-

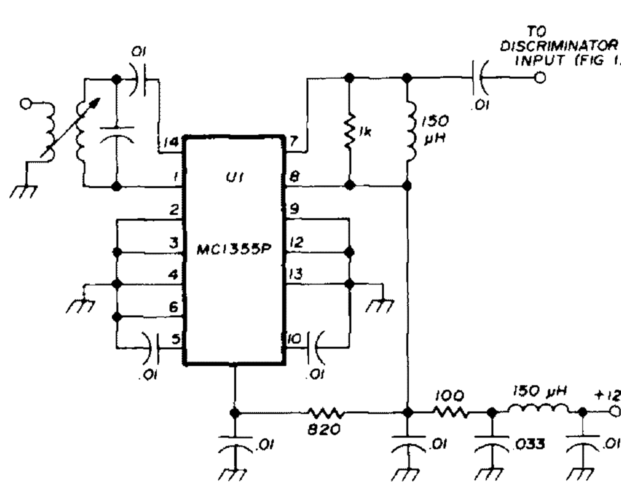

fig. 3. Amplifier/limiter for driving the crystal discriminator uses a Motorola MC1355P intergrated circuit.

covery from crystals that will not tune for a zero center. This just means that the crystal doesn't provide a good linear portion at exactly 9 MHz but with inexpensive, surplus crystals you can't be too particular.

Capacitors C1 and C2 are most easily adjusted with an audio-frequency sine wave applied to an fm signal generator or transmitter and using an oscilloscope to check distortion of the recovered audio sine wave. This same method should be used to align quadrature detectors, too, since there is no true zero-center reading. With a 1 volt p-p i-f signal (at 9 MHz) and 5 kHz deviation, the recovered audio will be about 1 volt p-p at the lower audio frequencies, rolled off at the higher frequencies with the de-emphasis network. The de-emphasis circuit also removes most of the remaining 9 MHz from the audio.

This type of crystal discriminator is amplitude sensitive and requires a good limiter ahead of it to provide a-m rejection. The slope of the discriminator is shown in fig. 2 but may vary a little with different crystals. It is possible to detect fm on the descending portion of the curve about 40 kHz lower but modulation acceptance is only a few kHz.

The Motorola MC1355P is an excellent IC to drive this discriminator. Fig. 3 shows a possible circuit, but the MC1355P has more than 60 dB gain available so it requires very careful layout, bypassing and shielding.

If you're not a reformed CBer it's possible to cultivate a friendship with a CBer who has recently acquired a vfo or you may opt to purchase a new crystal from JAN Crystals* for $2.50.

*JAN Crystals, 2400 Crystal Drive, Fort Myers, Florida 33901.

reference

1. Douglas A. Blakeslee, W1KLK, "Receiving FM, Part II," *QST*, February, 1971, page 16.

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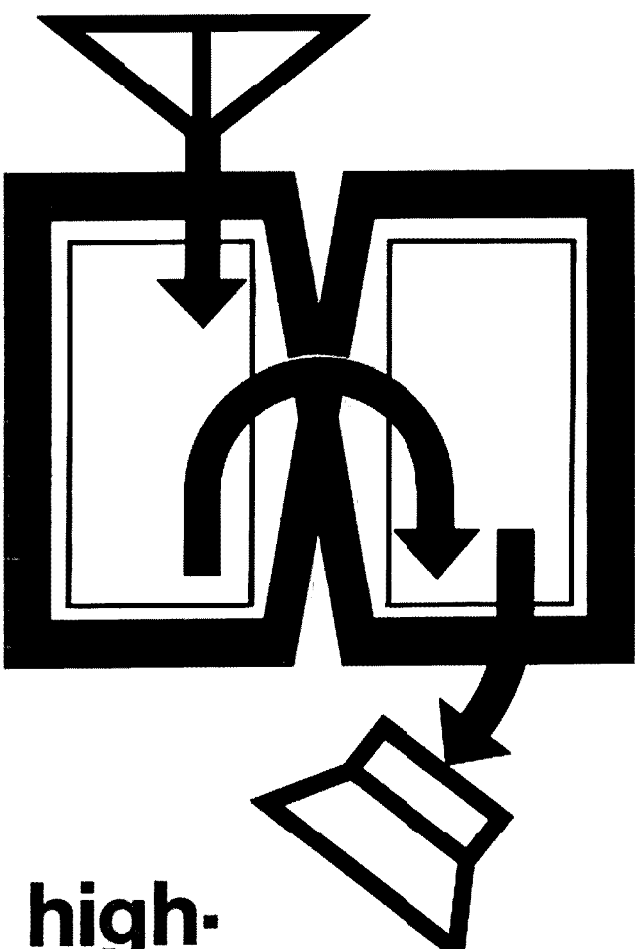

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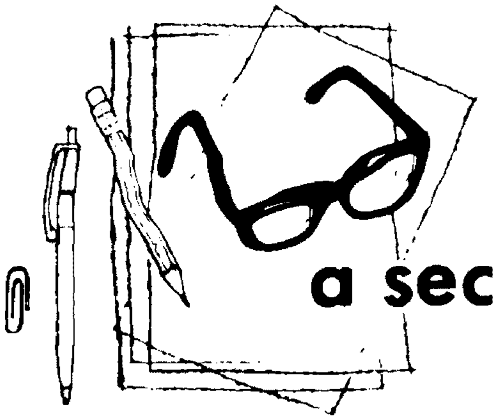


a second look

by jim fisk

Although MOS integrated circuits are finding widespread use in microprocessors, memories and other LSI (large-scale integration) applications, it appears that a relatively new form of bipolar logic, called I^2L (for integrated injection logic), can do everything its MOS rivals do – and probably better and cheaper. Although MOS manufacturers continue to squeeze more and more performance out of n-channel MOS technology, some researchers believe that the high performance-low cost characteristic of I^2L will end the dominance of MOS circuits in new generations of equipment.

Another characteristic of I^2L which intrigues designers is its versatility: although it doesn't directly lend itself to analog functions, it is compatible with bipolar manufacturing techniques used for linear devices so linear and I^2L can be combined on the same chip. Some digital-linear chips are already being developed, as are completely digital chips. In fact, according to one report, I^2L is now being designed into more circuit types by IC makers than are all MOS and other bipolar techniques combined!

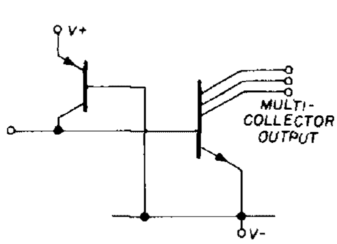
Originally formulated about four years ago at IBM's laboratories in Germany, and developed by Philips in the Netherlands, I^2L achieves MOS-level circuit densities by using planar npn transistors upside down (the basic logic element is an inverter). A direct result is the automatic isolation of all collectors, while the emitters are common. Lateral pnp transistors inject current directly into the base of a multi-emitter npn transistor operating in the inverse mode (see **fig. 1**). The result is a very simple gate structure which dissipates little power and has propagation delays on the order of 10 nanoseconds. This performance is comparable to that of standard TTL gates. By using integrated Schottky diode clamps speed can be pushed down to about 1 ns, making I^2L as fast as low-power Schottky TTL – Schottky-clamped I^2L, however, consumes 1/100th the power and is ten times smaller (circuit densities of 85 gates per square millimeter are routine).

Since I^2L units are powered through lateral pnp transistors, the circuitry is totally independent of resistors and can be operated over a wide speed range by simply varying the total current into the injector. Thus, the same I^2L device can run at slow speed in a watch, for example, dissipating microwatts, or at high speed in a microprocessor, dissipating milliwatts.

Many of the large semiconductor firms, including Fairchild, Motorola and Texas Instruments, are working on large-scale integration of I^2L and some devices are already on the market including TI's SBP0400 4-bit I^2L microprocessor. A 4096-bit I^2L random access memory may be available by the end of the year, and a 16-bit microprocessor with cycle times of less than 50 ns is expected sometime next year. No doubt MOS and TTL will be with us for a long time to come, but I^2L promises complex logic systems that could not be built economically with the older technology.

Jim Fisk, W1DTY
editor-in-chief

fig. 1. I^2L logic gate uses inverted transistor. Operation is completely independent of resistors and speed depends only on injected base current.



presstop

NEW CHIEF OF FCC'S AMATEUR AND CITIZENS DIVISION is John Johnston, K3BNS. The announcement was made during the FCC Forum at the ARRL National Convention in Reston, Virginia by FCC Safety and Special Services Bureau Chief, Charles Higginbotham, W3CAH. Charlie said the choice had been made official by the Commissioners only a few days earlier.

Johnston's Selection was a natural and will be widely welcomed by the Amateur fraternity. John had already established a fine track record with the Amateur and Citizens when he served there as Chief of the Rules and Legal Branch. He left Amateur and Citizens just a year ago to become Deputy Chief of the Spectrum Management Task Force. He originally joined the FCC in 1972.

"DE-REGULATION" will be the key word when Johnston picks up the reins at Amateur and Citizens. John plans to take a very hard look at the present rules to see where they can be relaxed to the benefit of both Amateur Radio and the Commission.

WARC 79 Working Group on Amateur Radio had its second full group meeting at Reston, with Prose Walker still in the Chairman's seat. Much of the all-day session was devoted to reports of the various task force chairmen, and it was obvious to the more than 30 attendees that the considerable effort that had already been invested was only a small part of the total job.

With Respect To Frequencies, the Working Group position is to strive for more spectrum in the HF bands both by making the bands we presently share with other services (and/or do not have at all in other parts of the world) exclusive worldwide Amateur bands, and by adding new Amateur bands. Proposed new HF bands would be 10.1-10.6 MHz, 18.1-18.6 MHz, and 24-24.5 MHz. It was also proposed that 40 meters be extended to 7.5 MHz, 20 meters to 14.5 MHz, and 15 meters to 21.5 MHz. At the low end of the spectrum a totally new band in the 150-200 kHz region will also be proposed. There is reason for hope that all or at least a good part of this expansion could be achieved, since some heavy users of the HF bands are moving to satellites; however, other services will be going for more HF frequencies, too. In the VHF/UHF spectrum, competition is tougher and the picture less clear — we'll have problems there.

INVERTED SPLITS for additional two-meter repeaters in the northern California area were selected as standard at "Sacramento '75." Northern California thus follows the lead of southern California, while the eastern seaboard goes the opposite way.

RIGHT-SIDE UP SPLITS were the choice of the Mid-Atlantic Repeater council at their meeting. Reasons were a wish to remain compatible with other East Coast areas, and encourage increased use of narrow-band gear.

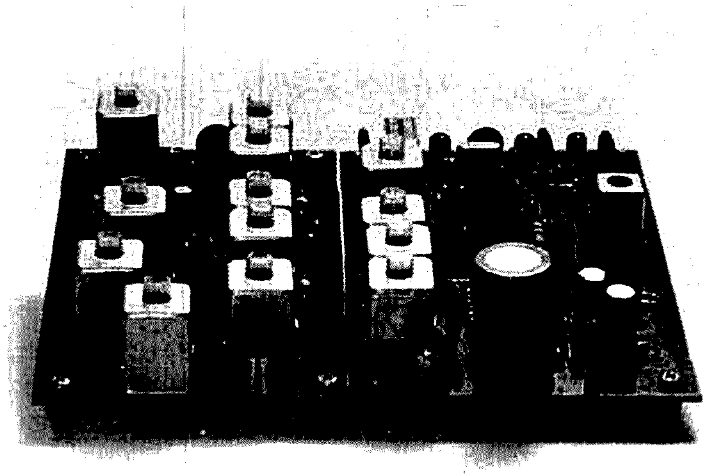
AMSAT'S EDUCATIONAL BULLETINS via OSCAR 6 resumed in September and will continue throughout the school year on mornings (U.S.) of even numbered days. Bulletin stations will transmit to be heard about 29.5 MHz on appropriate morning orbits as indicated by an "E" following the orbit number in the predictions.

Orbital Predictions from both HR Report and W6PAJ's booklet are both more than adequately accurate despite on-the-air comments to the contrary. Current HR Report sheet is within a few seconds, while W6PAJ's (prepared much earlier in the year) is accurate to within about a minute.

MULTI-2000, the multi-mode vhf rig which has caused interference in the aircraft band, is an offender primarily in its original version as imported by ITC, reports Mike Staal of KLM. The prime problem was with a spur ±16.9 MHz from the signal frequency and Mike reports that this has been corrected in the later versions which bear the KLM nameplate. All KLM Multi-2000s are being checked out with a spectrum analyzer to confirm that they meet published spurious specs.

All Early Multi-2000s should be checked out with proper instrumentation. Mike has some helpful suggestions for owners of the earlier radios — call him at KLM, (408)779-7363.

WORKING ALL STATES DURING 1976 will be rewarded by a very special Bicentennial WAS certificate from the ARRL. Only one award, for QSOs on any mode, any band — will be offered.



high-performance vhf fm receiver

Design and construction of a versatile fm receiver for use on any of the amateur bands from 28 through 220 MHz

Jerry Vogt, WA2GCF, Hamtronics, Inc., Rochester, New York

Are you looking for a compact, low-cost receiver to use with your new home-brew fm transmitter? Are you interested in trying fm without investing a lot of money right away for a transceiver? Do you need an extra fm receiver around the shack to monitor your local repeater or calling channel while you're operating on another frequency? This article describes a second-generation, solid-state vhf fm receiver which might be the answer. It is an improved version of an earlier receiver designed a few years ago[1] and uses two circuit boards: a vhf converter board and an i-f/audio board. The basic fm communications receiver may be used for 28, 50, 144 or 220 MHz (or adjacent commercial bands).

This new design includes the best features of its predecessor as well as refinements which improve selectivity and sensitivity, make construction and testing easier, and provide more flexibility. Built-in test points facilitate alignment and allow external signal strength and carrier frequency meters to be used.

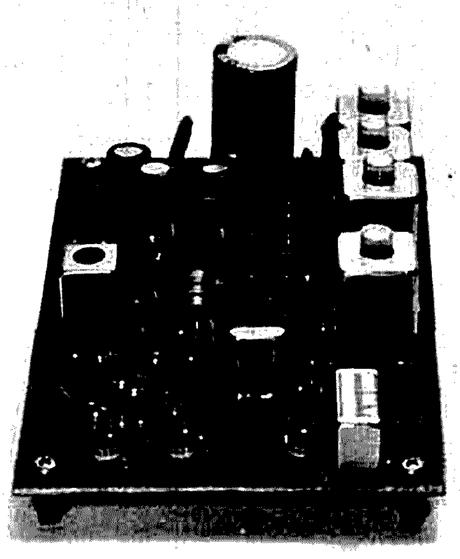
Stable, cascode circuits are easily tuned and require no neutralization. The sensitivity of the receiver is about 0.2 to 0.4 μV for 20 dB quieting. Adjacent channel selectivity is about 90 dB beyond the desired ±7.5 kHz passband. Image rejection is 40 dB. Operating power is 13.6 Vdc at 60 to 200 mA, depending on audio level.

Construction and alignment details are organized in three sections. The i-f/audio board is described first since it is straightforward and does not vary with the input frequency. Then the vhf converter board is described, along with variations for 10, 6, 2, and 1¼ meters. Finally, to demonstrate ideas for various receiver packages which can be based on the two basic boards, a short discussion of options is presented.

i-f and audio

The i-f/audio circuit (fig. 1) includes a sensitive and selective i-f amplifier, narrowband fm detector, audio amplifier and squelch circuitry. By including the proper external circuitry it may be used to build a single-channel vhf or uhf receiver, a multi-channel receiver, or a scanning receiver. The 10.7 MHz input to the board has a three-pole L-C filter operating into a low-noise grounded-base preamplifier, Q1. The high gain cascode mixer (Q2 and Q3) translates the input to a 455 kHz i-f with a selective ±7.5 kHz ceramic ladder filter providing 90 dB of adjacent channel selectivity and 40 dB image rejection. The 455 kHz i-f amplifier and limiter chain consists of five low-noise, high-gain transistor stages (Q5-Q9). The previous design used an IC — this discrete design provides improved operation, easier maintenance, and better metering.

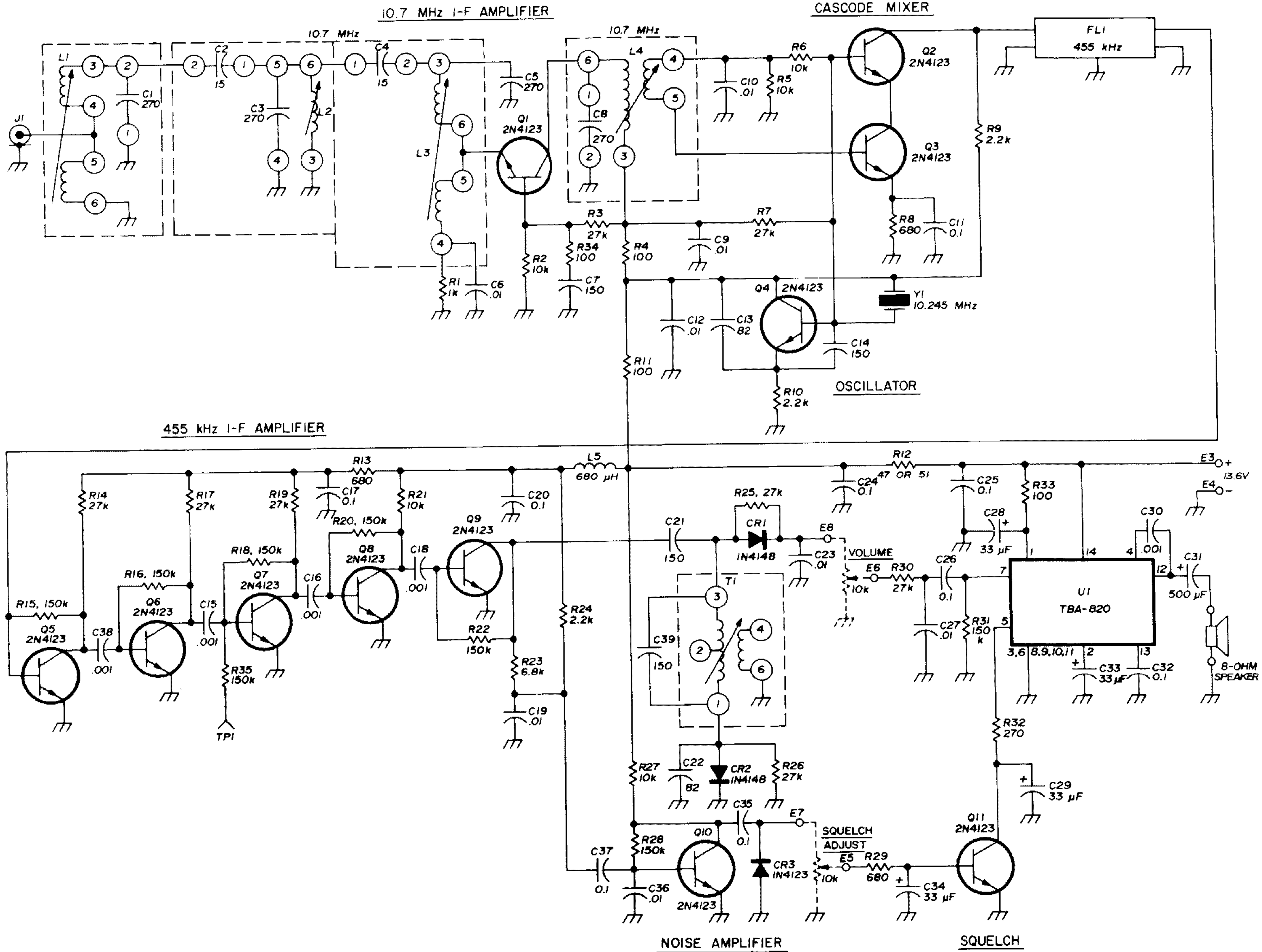
The i-f/audio board used in the vhf fm receiver. Murata 11-pole ceramic ladder filter is in lower left-hand corner.



The fm discriminator drives a special communications service 2-watt audio amplifier IC, an SGS ATE TBA-820, which incorporates lowpass filtering and de-emphasis to minimize hiss on weak signals. A sensitive squelch circuit (Q10-Q11) detects any a-m noise in the 7 kHz region to determine if a carrier is present in the limiter circuit. A flutter-proof circuit is used to prevent drop out of weak mobile stations. The audio circuit is set up so the user can change the volume control range or high-frequency response, if desired, to suit his own operating habits.

construction

Most pertinent construction details for the i-f/audio board are shown on the component location diagram (fig. 2). Following are details of coil assembly and other suggestions to facilitate proper assembly. The coils are wound on 10-32 (about 5mm diameter) plastic forms with carbonyl TH slugs. All are wound in a clockwise direction, as viewed from the top, using number 26 (0.4mm) solderable wire. All turns are close-spaced as shown in fig. 3. This drawing is exaggerated for clarity, but all leads should be pulled tight. No fancy bends are required, and no coil dope is necessary. Holes in the base of the form are numbered as indicated for


10.7 MHz I-F AMPLIFIER
CASCODE MIXER
10.7 MHz
10.7 MHz
FL1
455 kHz
J1
L1
L2
L3
L4
C1 270
C2 15
C3 270
C4 15
C5 270
C8 270
C10 .01
R5 10k
R6 10k
Q1 2N4123
Q2 2N4123
Q3 2N4123
R9 2.2k
R8 680
C11 0.1
R3 27k
R7 27k
R34 100
R4 100
C9 .01
R2 10k
C7 150
R1 1k
C6 .01
Q4 2N4123
Y1 10.245 MHz
C12 .01
C13 82
C14 150
R11 100
R10 2.2k
OSCILLATOR
455 kHz I-F AMPLIFIER
R13 680
L5 680 μH
R12 47 OR 51
E3 + 13.6V
E4 −
C24 0.1
C25 0.1
R33 100
C28 33 μF
C30 .001
R14 27k
R17 27k
R19 27k
C17 0.1
R21 10k
C20 0.1
R20, 150k
R18, 150k
R16, 150k
R15, 150k
Q5 2N4123
Q6 2N4123
Q7 2N4123
Q8 2N4123
Q9 2N4123
C38 .001
C15 .001
C16 .001
C18 .001
R35 150k
TP1
R22 150k
R23 6.8k
R24 2.2k
C19 .01
C21 150
R25, 27k
CR1 1N4148
E8
C23 .01
VOLUME
E6
10k
R30 27k
C26 0.1
C27 .01
R31 150 k
U1 TBA-820
C31 500 μF
8-OHM SPEAKER
C33 33 μF
C32 0.1
T1
C39 150
C22 82
CR2 1N4148
R26 27k
R27 10k
R28 150k
C35 0.1
E7
SQUELCH ADJUST
E5
R29 680
C34 33 μF
R32 270
C29 33 μF
Q11 2N4123
C37 0.1
C36 .01
Q10 2N4123
CR3 1N4123
NOISE AMPLIFIER
SQUELCH

reference when winding the coils. The coils can be prewound and then installed on the board with the keyways as shown. Primaries should be wound first, followed by the secondaries; then the capacitors, if any, are inserted through remaining holes in the base of the form.

Do not be overly concerned with coil winding. Neatness is not a requirement; the turns can overlap, and the windings don't have to be uniform. Secondary windings can be wound in a second layer over the primary or by continuing the first layer next to the primary. The only critical requirement is that primary and secondary of L1 and L3 must be correctly phased. (When the primary is finished at the tap, the secondary should start at the tap and be wound in the same [clockwise] direction.)

When the coil leads are inserted through the board, they should be started into the holes in the board while the coil form is spaced slightly away from the board; the form is then seated into place. Do not attempt to insert capacitor leads with the form tight against the board. After the coils are installed, application of heat from a very hot soldering iron for 10 to 15 seconds will automatically strip the wire. If you prefer, the leads may be stripped in the conventional way before installation. Do not solder-strip the leads unless the coil is mounted on the board as the leads will migrate into the plastic form.

Be careful, when installing the ceramic filter and the discriminator transformer, to seat them slowly by rocking to avoid lead stress. Resistor R35 is installed vertically with the top lead extending about 1/4 inch (6.5mm). to form a test point. Connections to the outside world are made by soldering number-22 (0.6mm) leads to pads on the board. The output circuit is designed for an 8-ohm speaker. However,

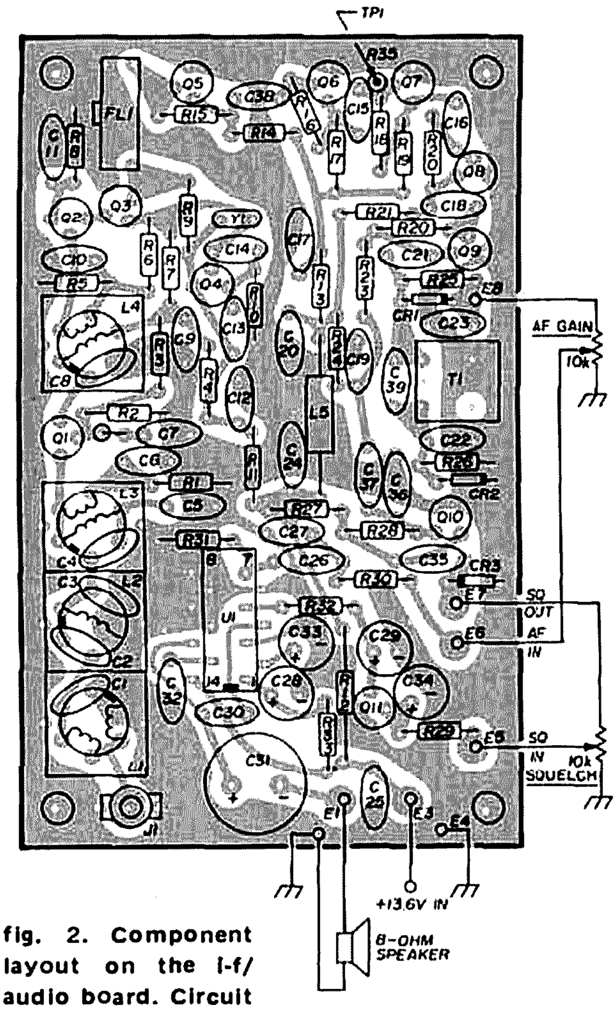


fig. 2. Component layout on the i-f/audio board. Circuit board is 2-3/4" wide (7cm) by 4½" (11.5cm) long.

fig. 1. I-f and audio system (left) for the vhf fm receiver has input at 10.7 MHz and includes five-stage 455-kHz i-f amplifier, squelch and two-watt IC audio amplifier. Filter FL1 is an 11-pole Murata CFS-455 ceramic ladder filter. Transformer T1 is a miniature 455-kHz i-f transformer with the internal resonating capacitor removed. The audio power IC is an SGS ATES TBA-820.

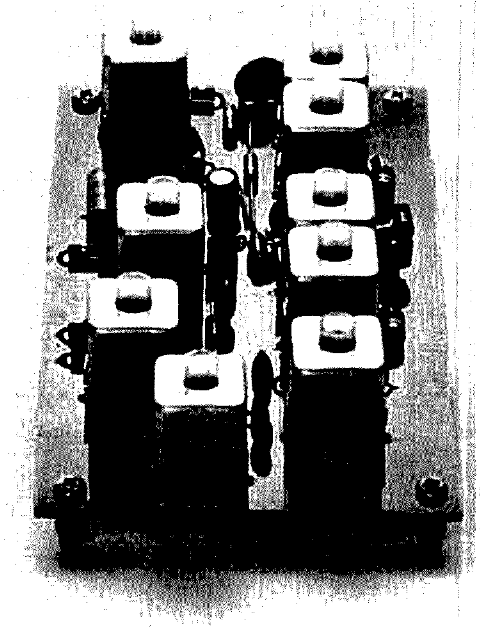
L1	Terminals 3 to 4: 10-5/6 turns no. 26 (0.4mm) on 10-32 (5mm) slug-tuned form; terminals 5 to 6: 2-5/6 turns no. 26 (0.4mm)
L2	12½ turns no. 26 (0.4mm) on 10-32 (5mm) slug-tuned form
L3	Terminals 3 to 6: 11½ turns no. 26 (0.4mm) on 10-32 (5mm) slug-tuned form; terminals 4 to 5: 3-1/6 turns no. 26 (0.4mm)
L4	Primary, 15½ turns no. 26 (0.4mm) on 10-32 (5mm) slug-tuned form; secondary, 9-1/6 turns no. 26 (0.4mm) on same form

other speakers may be used with some effect on frequency response and audio level.

alignment

With a 455-kHz signal generator connected through a dc blocking capacitor to the base of transistor Q5 and a vtvm connected to the top of the volume control (point E8), adjust transformer T1 for zero volt dc. Noise will be heard with no signal input, and the squelch should operate as expected. About 2 to 10 μV at 455 kHz should provide 20 dB quieting. Now set the signal generator to 10.7 MHz and couple it to the input, J1. Alternately peak L1 through L4 for maximum negative voltage at TP1 (top of R35). Image response at 9.79 MHz should be down about 40 dB, and the sensitivity at 10.7 MHz should be 2 to 10 μV.

If you wish to use meters to indicate signal strength or carrier frequency, this may be done. A zero-center 50 μA meter may be connected in series with the top of the 10k volume control (connect a small electrolytic capacitor across the meter). The dc current through the volume control will operate the meter, with positive swings indicating high-frequency error and vice versa. A sensitive voltmeter circuit can be built around a Darlington pair or an op amp to drive an S-meter from the limiter test point at R35. Do not load the base of Q7 by changing the value of R35 or by making connections directly to the base of the transistor. The voltage swings from +0.6 volt to about −3 volts, so some bias is required in the meter amplifier to avoid swinging through zero on the meter.



Vhf converter board for the fm receiver. Although designed for use with the 10.7 MHz i-f/audio board, it can be used with other i-f systems as discussed in the text.

Since some operators may prefer different frequency response or volume control range, the following information is provided as a guide. The value of R32 may be reduced as low as 47 or 51 ohms to increase audio gain. A corresponding increase in the high frequency response also results. The value of C27 may be changed to vary the audio frequency response. A 0.1 or 0.05 μF capacitor here will provide bass response or more de-emphasis; a 0.001 μF capacitor at C27 will increase high-frequency response.

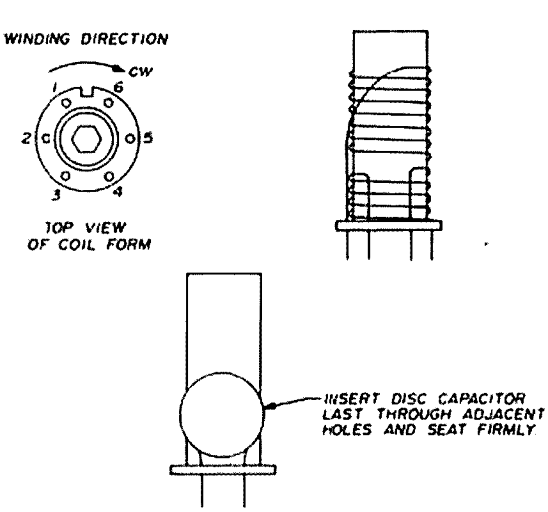


fig. 3. Coil winding and lead identification.

vhf converter

The vhf converter consists of a sensitive cascode rf amplifier, low-noise fet mixer, an oscillator, and an injection multiplier/buffer chain. Two schematic

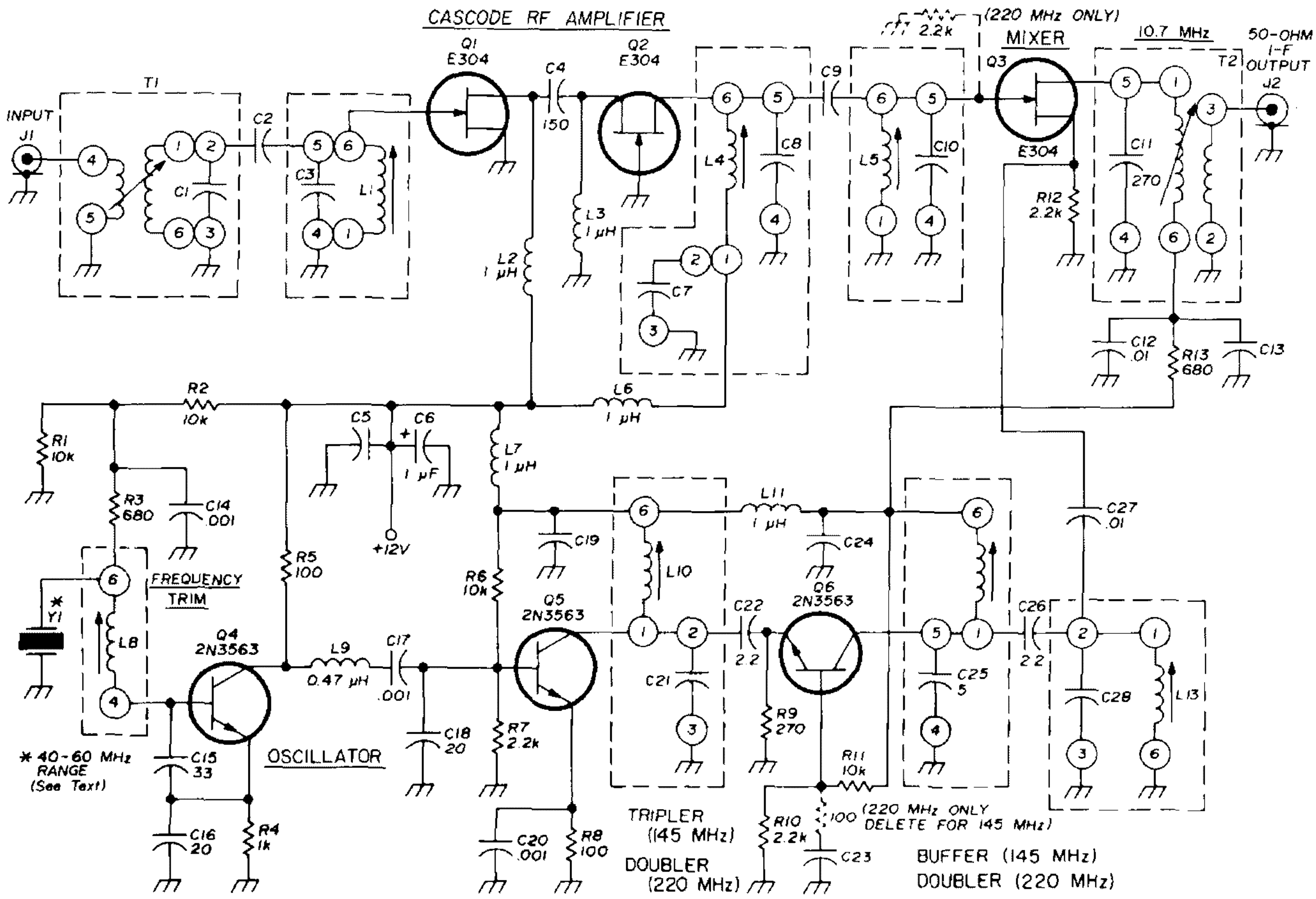


fig. 4. Schematic of vhf converter designed for use on two meters (values shown are for 145 to 155 MHz) and 220 MHz. Inductors and transformers are wound with no. 26 (0.4mm) wire. Tuned-circuit values *may* be changed as required for operation in adjacent commercial bands.

	145 MHz	220 MHz
C1	10 pF	3.9 pF
C2	1.0 pF	0.68 pF
C3	15 pF	5 pF
C5	1000 pF	82 pF
C7	150 pF	82 pF
C8	5 pF	3.9 pF
C9	1.0 pF	0.68 pF
C10	15 pF	5 pF
C13	270 pF	82 pF
C19	150 pF	600 pF
C21	15 pF	20 pF
C23	270 pF	150 pF
C24	150 pF	270 pF
C28	10 pF	20 pF
L1	2-1/6 turns	2-1/6 turns
L4	4-5/6 turns	3-1/6 turns
L5	2-1/6 turns	2-1/6 turns
L8	14-1/3 turns	14-1/3 turns
L10	2-5/6 turns	3-5/6 turns
L12	4-1/6 turns	2-1/6 turns
L13	2-1/6 turns	2-1/6 turns
T1	Primary, 1-1/6 turns; secondary, 3-1/6 turns	Primary, 1-1/6 turns; secondary, 3-1/6 turns
T2	Primary, 14-1/6 turns; secondary, 2-5/6 turns	Primary, 14-1/6 turns; secondary, 2-5/6 turns

diagrams (**fig. 4** and **5**) give details for various bands, including tuned circuit variations, bypass values, and local-oscillator chain.

The i-f output normally is 10.7 MHz for use with the i-f/audio board. However, the output transformer can be modified to cover other intermediate frequencies, such as 14, 28, or 50 MHz, if you wish to use the converter with a tunable receiver as an i-f.

The converter board includes one oscillator. Multichannel operation may be accomplished by using a multichannel adapter[2] in place of the built-in oscillator. The converter may be used for scanner operation by switching oscillator frequencies in the multichannel adapter.

The crystal in the converter is a third overtone, 0.002% unit cut for series resonance less 1000 Hz (many two-meter transceiver crystals may be used).

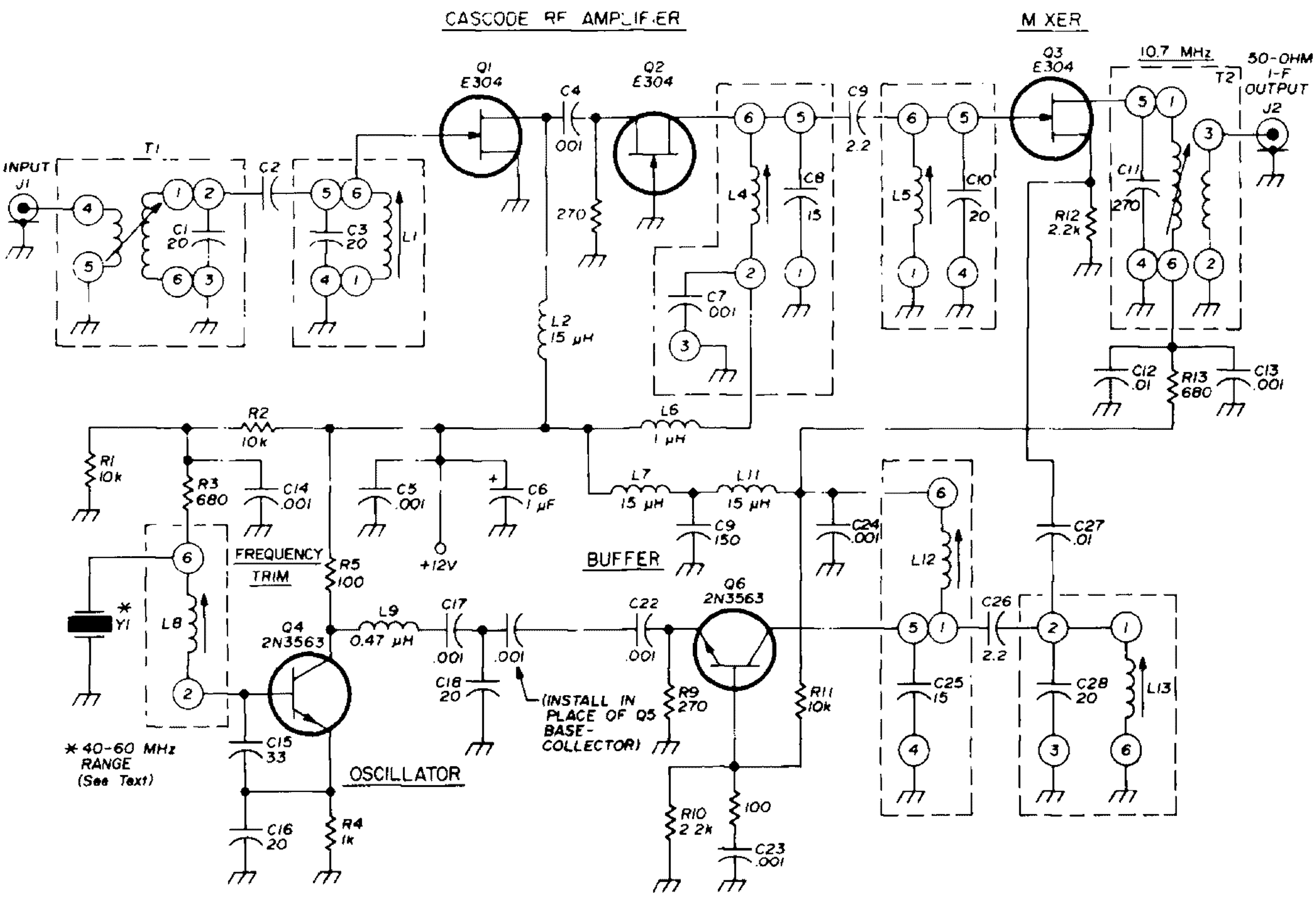
The required crystal frequency is given by

$$\frac{\text{channel frequency} - \text{i-f}}{X}$$

where X is the frequency multiplier. For channel frequencies of 30 to 60 MHz, X be trimmed to the desired operating frequency.

construction

Most pertinent construction details are shown on the component location



	28 MHz	50 MHz
L1	16-1/6 turns	9-1/6 turns
L4	17-1/6 turns	10-1/6 turns
L5	16-1/6 turns	9-1/6 turns
L8	14-1/3 turns	14-1/3 turns
L12	13-1/6 turns	13-1/6 turns
L13	8-1/6 turns	8-1/6 turns
T1	Primary, 5-1/6 turns; secondary, 20-1/6 turns	Primary, 5-1/6 turns; secondary, 11-1/6 turns
T2	Primary, 14-1/6 turns; secondary, 2-5/6 turns	Primary, 14-1/6 turns; secondary, 2-5/6 turns

fig. 5. Vhf converter for use on the six- and ten-meter amateur bands. All inductors and transformers wound with no. 26 (0.4mm) wire.

= 1; 90 to 130 MHz, X = 2; 130 to 180 MHz, X = 3; and 180 to 230 MHz; X = 4. The crystal should be cut about 1000 Hz less than the calculated frequency. This fudge factor allows the crystal to

diagram, **fig. 6**. Coil forms are the same as used in the i-f/audio board, except that carbonyl J slugs are used in the coil forms. Winding information for the i-f coil is for 10.7 MHz. For an i-f near 14

MHz, the primary winding should be reduced to about 10-1/6 turns. For 28 MHz, the primary should be 7-1/6 turns, and the secondary should be 1-5/6 turns. For 50 MHz, capacitor C11 should be changed to 15 pF, and the turns should be as shown on the schematic.

As a matter of interest, the unconventional long leads on a few of the components and the +13 volt connection to the center of the board permit maximum ground area in the board layout. In effect, you get the ground plane performance of a double-sided circuit board without the problems encountered in working with two foils.

When building the converter be sure to observe polarity on the electrolytic capacitor, and be sure to solder the shield can lugs to the board. If coil pruning becomes necessary, the shield cans may be unsoldered. All components should be seated close to the board to provide short leads. If a multichannel adapter is to be used, the oscillator on the converter board can be included for test purposes and later disabled when the adapter is connected.

Phono connectors are used to allow easy connection to the board with coaxial cable. This may be done at a tuned circuit because the coax is terminated at such a point. However, any connectors used in mid-line should be constant-impedance types for low loss, and phono and type-uhf connectors may put a bump in the line in such applications. Likewise, the cable should be chosen carefully for low signal levels. RG-8/U cable (or better) should be used unless you can accept the higher loss of the smaller cable types. If a separate transmitter is used with the converter, a good coax relay should be used to minimize signal loss and to prevent coupling of large amounts of rf into the front end of the converter.

converter alignment

The most difficult part of the alignment procedure is obtaining a stable test signal. Even my HP-608 signal generator takes several hours to settle down enough to stay within a 5 kHz passband at vhf. An alternative is a crystal-controlled weak-signal source such as

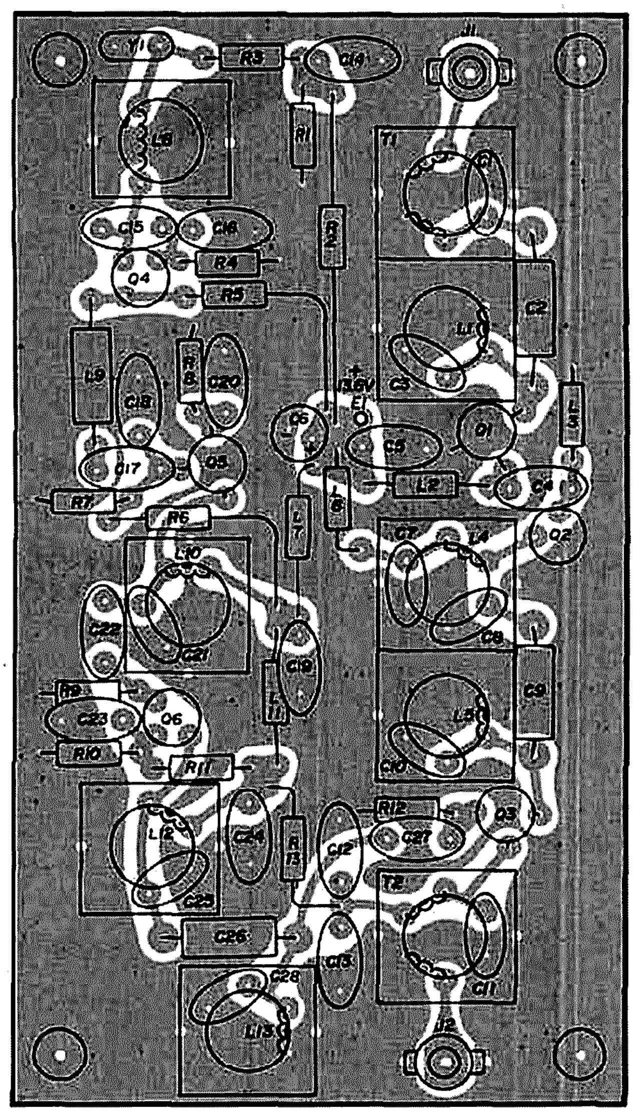


fig. 6. Component layout for the vhf converters. Same circuit board is used for each of the converters shown in figs. 4, 5 and 6. Circuit board is 2½" (6.5cm) wide and 4½" (11.5cm) long.

those which have been described in the past. An on-the-air test, if it can be arranged, is another possibility.

Start with all adjustments at about half range. Tune in a signal, and peak all adjustments. If the coils do not peak within the range of the slug, an adjust-

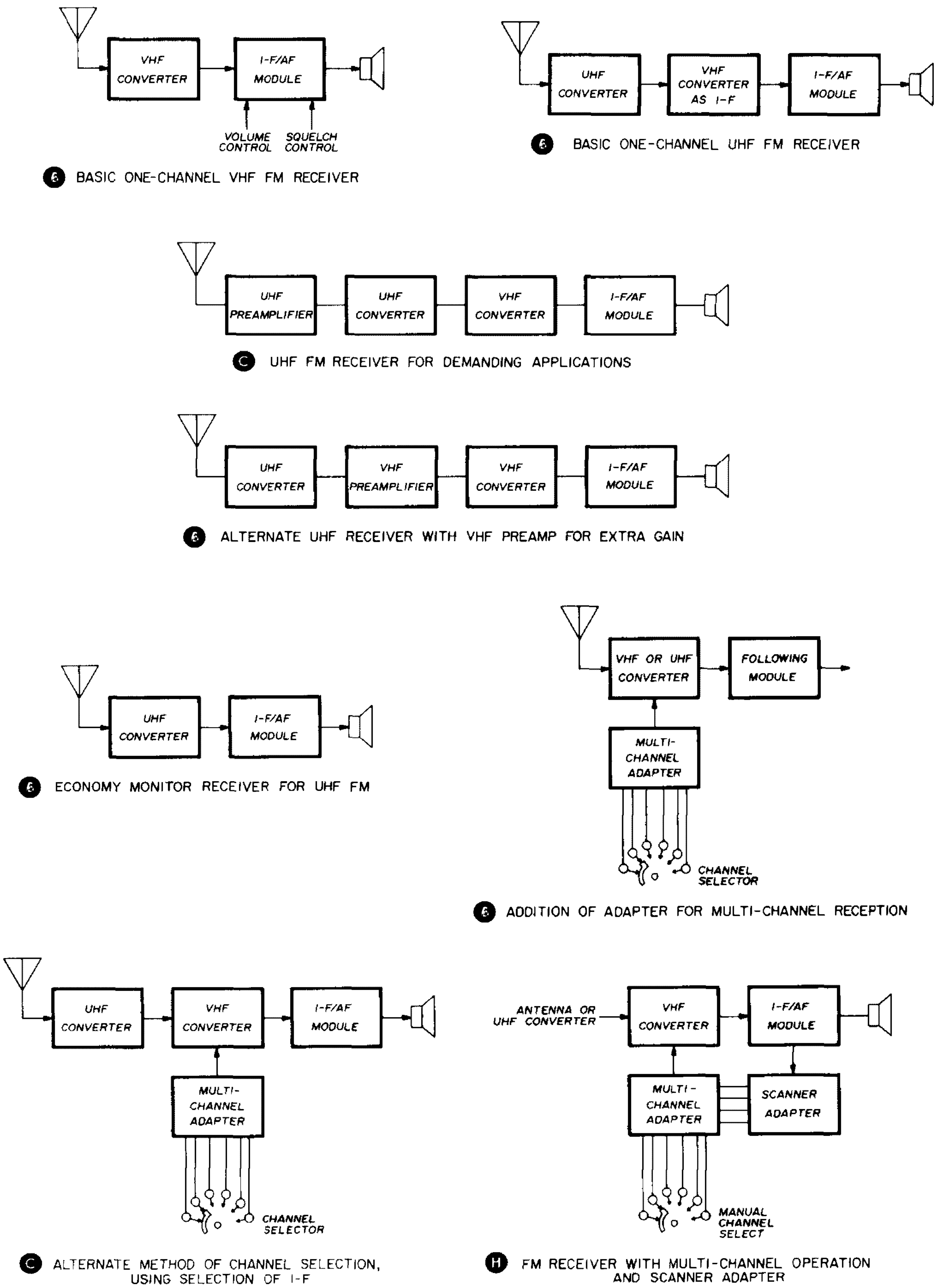


fig. 7. Some ideas for complete vhf receiver systems using the vhf converter and i-f/audio board described in this article. Kits for each of the circuits shown here are available from Hamtronics.*

ment in the number of primary turns may be necessary. Be careful, however, that you don't tune a multiplier coil to the wrong harmonic. Then, adjust the oscillator trimmer coil (L8) to net the converter to the channel frequency by monitoring the receiver discriminator or S-meter. Note that the crystal may be pulled enough for adjustment over a range of about 4 kHz at vhf. A vtvm connected to test point TP1 may be used for peaking adjustments when aligning a converter which will be used with the previously described i-f/audio board.

The final alignment should be done by peaking all rf, i-f and multiplier or injection coils with a weak received signal. Antenna reactance may require that the input coil be repeaked when the antenna is connected. Because of interactions between pairs of coils, such coils should be peaked alternately until you find the combination which provides the test sensitivity. This is especially true of L12 and L13, which are somewhat overcoupled. There should not be any tendency to oscillate when the coils are peaked.

When used with the i-f/audio board, the converter should provide sensitivity of about 0.2 to 0.4 μV for 20 dB quieting. Meter action at TP1 on the i-f/audio board should start with as little as 20 μV of signal into the converter.

If a multichannel oscillator is used in place of the converter's local oscillator, R5 and Q4 should be removed from the converter. The following parts also may be removed if desired: R1, R2, R3, R4, C14, C15, C16, Y1, L8 and shield.

receiver system ideas

After building the basic receiver you may wish to add accessories to extend its usefulness. **Fig. 7** illustrates a variety of receiver configurations using the two boards described in this article as well as circuit boards featured in earlier articles.

The arrangement *in* **fig. 7A** is the basic setup described in this article. The layout in **fig. 7B** uses the uhf converter described in a previous article[2] for coverage of the 450-MHz amateur band. For weak signal uhf reception or long-distance communications, a uhf preamplifier may be included as shown in fig. 7C. An alternate layout that provices good uhf performance is shown in **fig. 7D.**[3,4] For uhf monitor service, the simple circuit of **fig. 7E** is recommended.

Fig. 7F shows how a multi-channel adapter may be added to the circuit for multi-channel operation. A multi-channel fm receiver with a scanner adapter[5] is shown in **fig. 7H.**

*The following kits are being made available in conjunction with this article. Be sure to specify exactly what you want, including frequency band.

I-f/Audio Board kit	R40	$40.00
Vhf Converter kit	C25	25.00
Receiver kit (both of above)	R60	64.95
Vhf Preamplifier kit	P6	6.00
Six-Channel Adapter kit	A13-45	12.95
Scanner Adapter kit	AS-10	10.00
Uhf Converter kit	U20-450	20.00
Uhf Preamplifier kit	P15-450	15.00

When ordering please add shipping; New York residents please add sales tax. Quantity prices are available to clubs and to individuals who are interested in distribution at hamfests, etc. A complete catalog is available in exchange for a self-addressed, stamped envelope. Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

references

1. Gerald Vogt, WA2GCF, "VHF FM Receiver," *ham radio*, November, 1972, page 6.
2. Gerald Vogt, WA2GCF, "Uhf Converter and Preamplifier," *ham radio*, July, 1975, page 40.
3. Gerald Vogt, WA2GCF, "Improved 6-Meter Preamplifier," *ham radio*, January, 1973, page 46.
4. Gerald Vogt, WA2GCF, "Improved 2-Meter Preamplifier," *ham radio*, March, 1972, page 25.
5. Gerald Vogt, WA2GCF, "Channel Scanner for VHF FM," *ham radio*, November, 1974, page 26.

ham radio

using TTL ICs in single-sideband equipment

Simple TTL IC ssb circuits include a complete transceiver using only three SN7400 NAND gates

Peter J. Hampton, G4ADJ

A while ago I was toying with TTL crystal oscillators for use as clocks in digital equipment. The performance of standard multivibrator configurations (see fig. 1) was quite surprising. Just about every crystal I had performed equally well in the circuit. Fundamentals from 100 kHz to over 20 MHz all gave outputs of at least 2 volts peak-to-peak. Perhaps, I thought, these circuits could be used successfully as local oscillators in high-frequency ssb equipment.

Since the output of the circuit of fig. 1 is a square wave, lower second harmonic content can be expected as compared to most conventional oscillators.[1] This can be a very good thing when trying to filter out the spurious responses so often troublesome in homebrew ssb gear.

mixer

Having found a cheap, sure-fire local oscillator, A TTL-compatible mixer was required to provide the appropriate double-sideband signal. For this purpose nothing more complicated than a single NAND gate was found to be necessary. Two square waves, $f1$ and $f2$, when applied to the separate inputs of a NAND gate, yield outputs of $f1 + f2$ and $f1 - f2$. Hence the gate is performing the function of a *product modulator*. If, however, one input to the gate is biased at the point (A) on the transfer characteristic shown in **fig. 2**, then a small signal applied to that input will be amplified linearly (see reference 2), and

also switched by the square-wave signal at the other input. Thus, if the switching signal is an rf carrier and the other a speech waveform, then the gate becomes a low level amplitude modulator. A carbon or crystal microphone used as the audio source will usually provide enough output to give a 100 per cent modulated signal.

Reference 2 states that a TTL gate can give large amounts of gain at frequencies as high as 10 MHz when operating as a linear amplifier. Thus it was decided to see if the device could be used as a simple and inexpensive rf

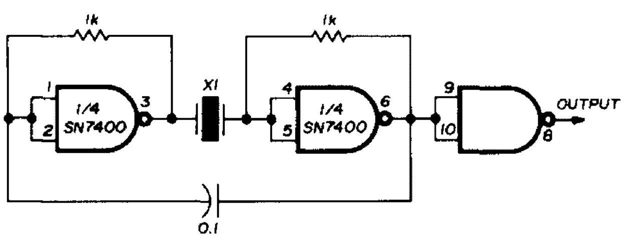


fig. 1. TTL oscillator for fundamental crystals operating in the range from 100 kHz to 20 MHz.

amplifier for small signals. The circuit of **fig.** 4 was lashed up, and, to my surprise, it performed very well indeed when correctly biased. A gain of around 15 dB was obtained at 8 MHz and there seemed to be no major instability problems. However, it is recommended that double-sided PC board be used with one side acting as a ground plane. Also 0.1 μF and 100 μF capacitors should be wired across the supply pins of each IC to provide adequate decoupling of the +5 volt bus.

ssb exciter using TTL gates

In the junkbox at home I found a large number of 10XJ crystals of surplus origin with fundamentals of 3.446 and 3.449 MHz. Not being able to think of anything else to do with them, I attempted to make up something of an ssb filter for use in an exciter built around circuits similar to those outlined

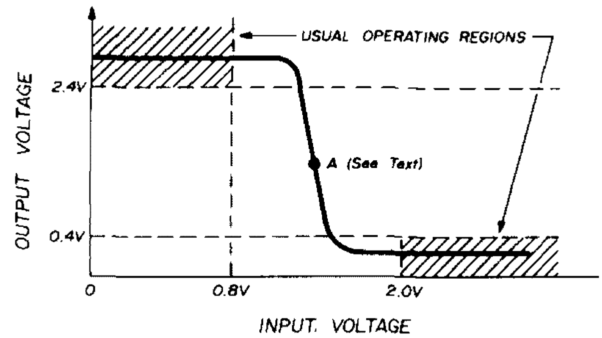


fig. 2. Transfer characteristics of the TTL NAND gate.

above. The unit was designed to generate and detect upper-sideband signals at 3.447 MHz. **Fig.** 5 shows the circuit of the complete prime-mover.

The oscillator section is straight forward enough and uses a 3.446 MHz crystal for upper sideband and 3.449 MHz for CW. Gates U1C and U1D are buffers for transmit and receive, respectively. These are followed by controlled gates U3B and U2A which route the carrier to the appropriate mixer while shutting the unused one off. PTT (or full break-in CW) is provided by U3A and U2B.

Gates U2C and U3C are the modulator and detector, each giving an output of 2 volts p-p for inputs of 100 mV or so. Following the modulator is a filter/amplifier arrangement comprised of crystals X2, X3, X4 and U2D which is used to supply a certain amount of rf clipping before the main filter. U3D, the remaining gate, is used as an rf amplifier preceding the product detector U3C.

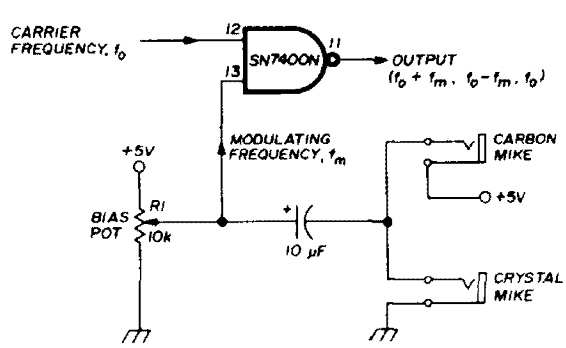


fig. 3. Using the SN7400 gate as a product modulator (mixer).

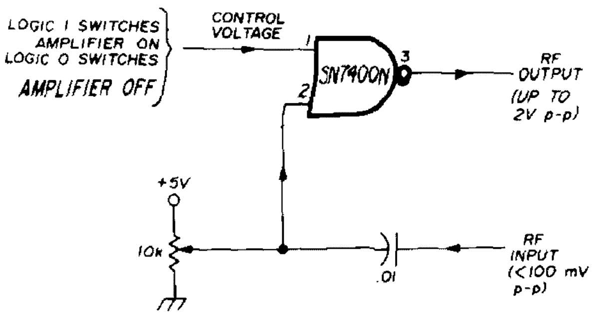


fig. 4. Simple rf amplifier using a TTL NAND gate. Rf output is about 2 volts p-p for 100 mV p-p input.

Both U3D and U2D are controlled by the PTT voltage applied to their unused inputs.

the filter

The *pièce de résistance* of most ssb rigs is their filter. Well, this one has quite a job to do since it must remove all the carrier from an a-m signal (the modulator is unbalanced) and also have a reasonable passband characteristic for good audio reproduction. The simple ladder arrangement shown in **fig. 6** seemed to work pretty well, but with an insertion loss of around 10 to 12 dB.

All the series elements of the filter use 3.449 MHz crystals with additional capacitance shunted across them to give series resonant frequencies spread throughout the range from 3.4465 to 3.4485 MHz. The remaining shunt crystals are all resonant at 3.4460 MHz, the carrier frequency, and effectively shunt it to ground.

In the original a total of twelve 10XJ crystals were used, giving about 45 dB of carrier and 35 dB of lower-sideband suppression. There is plenty of room for improvements in the filter design, however!

summary

In conclusion, we have here the basis of a very inexpensive unit which can generate and detect ssb signals at good quality with a minimum of external components. Transmit output power is

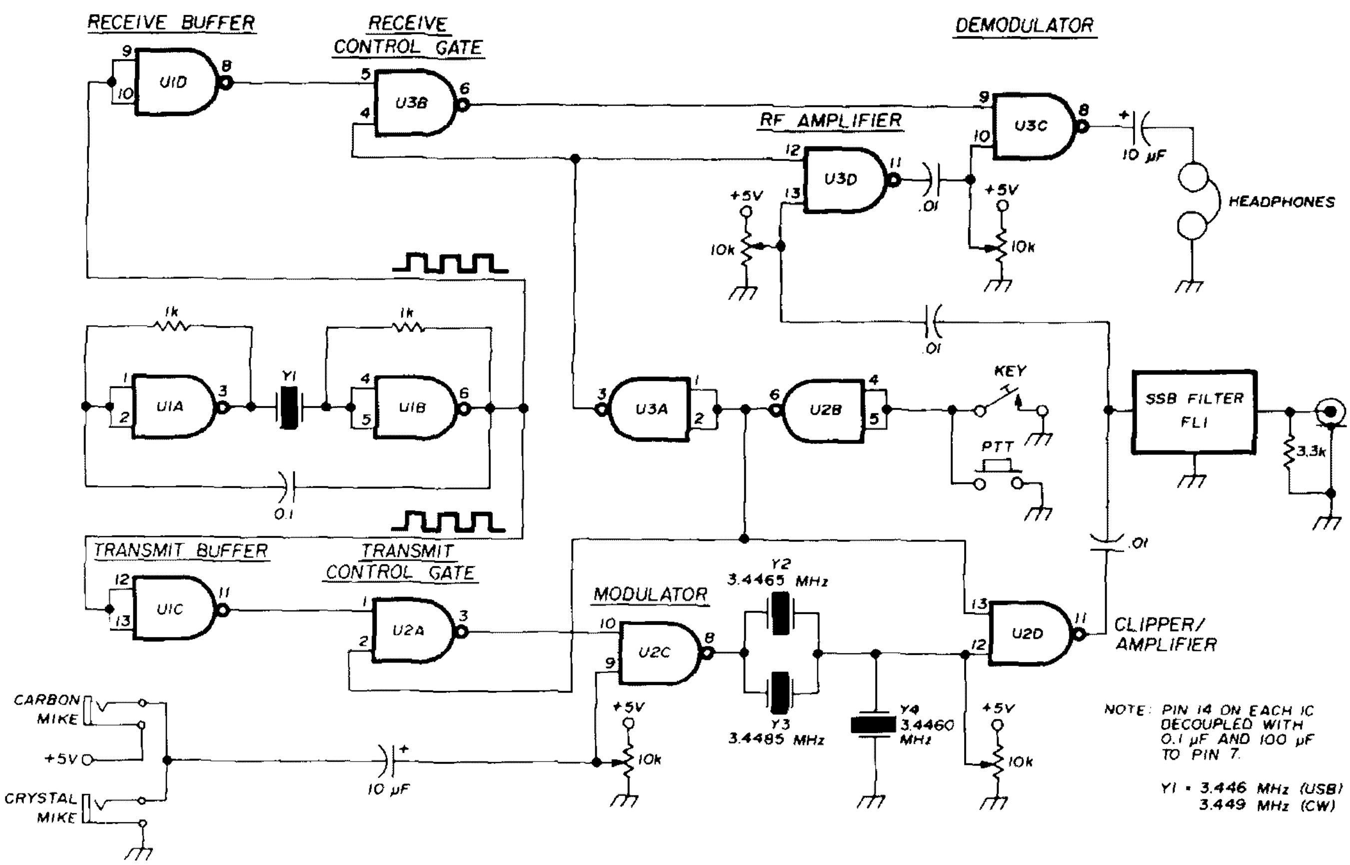


fig. 5. Schematic diagram of a complete ssb transceiver using 7400 series ICs. Circuit for ssb filter, FL1, is shown in fig. 6.

on the order of 10 mW PEP and receiver sensitivity is about 50 mV for noise-free audio (i.e., an S9 input signal). It is intended that the rest of the receiver gain be supplied by the preselector stages in the frequency translator.

is a good idea to use an oscilloscope for setting the bias points as there is a considerable variation in gate characteristics between samples. This is why potentiometers are shown in all the previous circuits instead of fixed resistors.

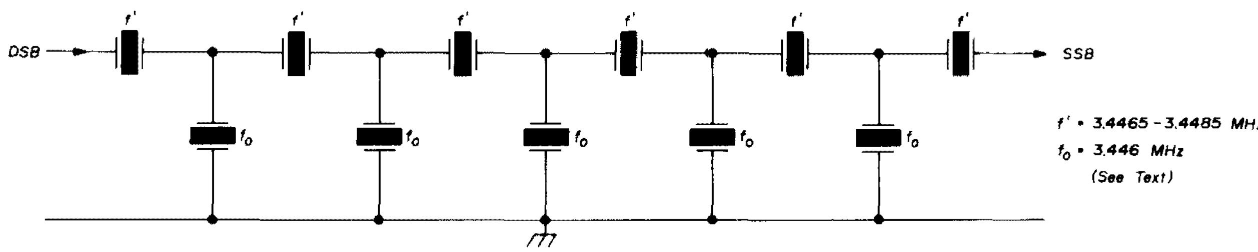


fig. 6. Simple ladder filter for 3.447 MHz upper sideband using surplus 10XJ crystals.

As far as I can see, the standard TTL gate is probably the most useful active device available to the home constructer on a cost/performance basis. It seems that it can be used in almost any application up to 20 MHz or so with a minimum of external components and little difficulty in setting up. Nevertheless, it

references

1. Max Robinson, K4ODS, and John Smith, "Local Oscillator Waveform Effects on Spurious Mixer Responses," *ham radio*, June, 1974, page 44.
2. *Texas Instruments System 74 Designer's Manual*, Texas Instruments, Inc., Dallas, Texas, page 20, note iii.

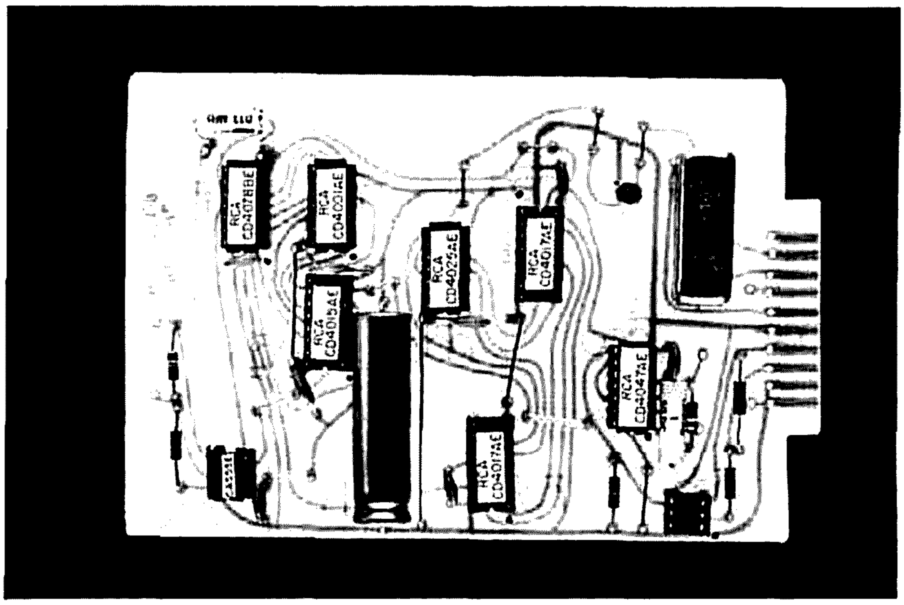
ham radio

Kenwood TS-520 CW filter option modification

Owners of the Kenwood TS-520 transceiver who have the CW filter installed are confronted with the problem that all CW reception must be with the narrow filter when transmitting in this mode. Often it's desirable and more pleasant to receive CW with a wider bandpass as on the upper and lower sideband modes. A very simple modification consisting of the installation of an auxiliary switch permits the option of the wideband or CW narrowband filter. Fortunately the TS-520 mode control employs diode switching when inserting the CW filter. Thus lead length and capacitance are no problem, and the CW filter control leads can be extended and switched remotely or externally.

The TS-520 has a flat plate on the chassis underside, which is part of the dial assembly and which is an ideal location for such a switch. Adequate clearance is available to install a miniature spdt switch by carefully drilling a hole in this plate for mounting the switch and a clearance hole in the bottom of the outer cover case to permit the switch handle to protrude without being obvious or defacing any panel space. A small three-wire cable connects the switch to the filter control circuit by attaching the center pole of the switch to the original brown common wire and connnecting one switch pole to the CW terminal on the TS-520 i-f circuit board and the other switch pole to the ssb terminal. Thus you now have the option of a wideband CW position (ssb filter) or the 500-Hz CW filter by the simple flick of this switch.

Bill Vandermay, W7ZZ



RTTY line-end indicator

Solid-state RTTY line-end indicator uses CMOS logic ICs for high reliability and low current drain

Robert Mendelson, W2OKO, RCA Solid-State Division, Somerville, N.J. 08876

Can you type a smooth 60 words per minute? Would you like to have your RTTY transmissions sound like commercial press sent from a tape? It really is quite easy to do provided you are equipped with a tape perforator and transmitter-distributor.

Unfortunately, many amateurs have the equipment but don't like to punch tape while receiving the other fellow's message. This is especially true when there is only one keyboard and printer in the station. You often hear, "How can I punch tape without seeing what I'm typing; how will I know when I am near the end of the line?" Or, "I have two machines, but if I run them both at once my wife would throw me out of the house!" Relax! It can all be done with one machine with the printer copying the incoming traffic while the keyboard types the answers. Typing blind is really not difficult, and an occasional error will not really matter during a ragchew. It's the end of line that is annoying.

Articles have been written about RTTY line-end indicators,[1] but they have all been based on mechanical switches, or counting word spaces instead of letters, or other similar circuits. With today's digital and linear ICs it can all be done electronically with the line-end light or bell actuated at 66 characters every time.

The circuit shown in fig. 1, which uses RCA CMOS digital ICs and a bipolar timer, does this. The CMOS ICs have many advantages over the more familiar TTL logic family. Power consumption is minimal. The whole circuit draws 5 mA quiescent or operating,

except when the light goes on. The power supply voltage can be anything from 4.5 to 15 volts and does not really have to be regulated.* Further, the circuit needs only a few common resistors and capacitors.

circuit operation

In the circuit of fig. 1 an optical isolator, U1, in the 60 mA loop will turn its transistor output on and off in response to the mark-space code. The output is separated completely from the input and allows the 120-volt loop supply to be applied without any danger to the CMOS circuits. The zener diode across the input protects the optical isolator against incorrect polarity or too high a source voltage.

The isolator collector is tied to the input of U2, an RCA CD4047 monostable oscillator. The appearance of a start pulse will trigger the oscillator whose holding time is set by R1 and C1 to 150 milliseconds, the time it takes for the start pulse, 5 code pulses, and part of the stop pulse to occur. The isolator output is also fed to U5, an RCA CD4015 shift register. This function will be explained later. Each time a character fires U2, its output will trigger U3, an RCA CD4017 divide-by-10 counter which, after 10 pulses, will trigger U4, another RCA CD4017.

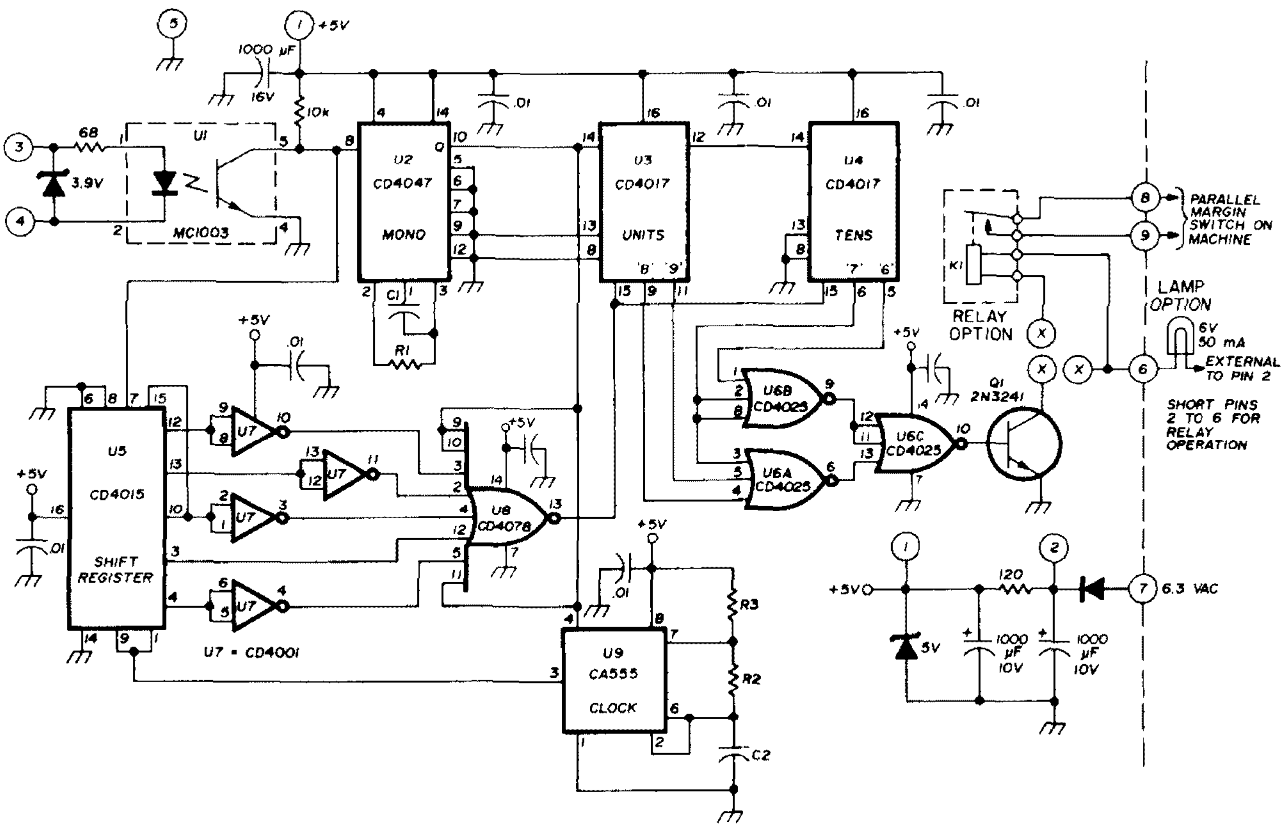


fig. 1. Schematic diagram of the RTTY line-end indicator. Correct values for C1, C2, R1, R2, and R3 are shown for speeds of 60 and 100 wpm. Power supply option for 6.3 Vac input is shown at lower right. Circled numbers refer to PC connector pins shown in fig. 2.

speed	C1	C2	R1	R2	R3
60 wpm	0.1 µF	0.033 µF	560k	470k	56k
100 wpm	0.1 µF	0.033 µF	330k	270k	56k

The CD4017 has a serial input and output, but it also has ten outputs, one for each digit. Only one of these outputs will be at logic 1, or high, at any time. Each input pulse will move the

*A zener regulator will help hold the clock frequency if the lamp load should cause a large change in supply voltage.

logical 1 state from pin to pin until, on the tenth pulse, it will be back where it started.

This rotating logic 1 can be used to form a sort of combination lock when used in conjunction with some simple NOR gates. The output of a NOR gate will go high only when all of the gate inputs are low. Using this information, let's work back from the lamp which will signal that the end of the RTTY line is near.

The lamp will be on when Q1 conducts; that is, its base is greater than 0.7 volt. The base voltage is supplied by the output of U6C, a NOR gate. For this gate to have a high output, its inputs from sections U6A and U6B must be low. For a count below 60 the leads from 6 and 7 of the tens counter will both be low, causing the output of U6B to be high. Under these conditions the output from U6C cannot go high to turn the lamp on.

At count 60 the output of section U6B will go low, but from count 60 to 67, the outputs from U3 units counts 8 and 9 will be low, as will the connection from the U4 7 count. Therefore, the output from U6A will be high and the output from U6C will stay low -- the lamp stays off.

However, at count 68, the units 8 will go high, driving the output of U6A low. Since the U6B output is also low, output from U6C will go high and the lamp will turn on. This same condition will exist at count 69. At count 70 both U6A and U6B will have a high input (from the 7 count on U4) and the lamp will stay on until count 80, which is way past the line end of 72 characters.

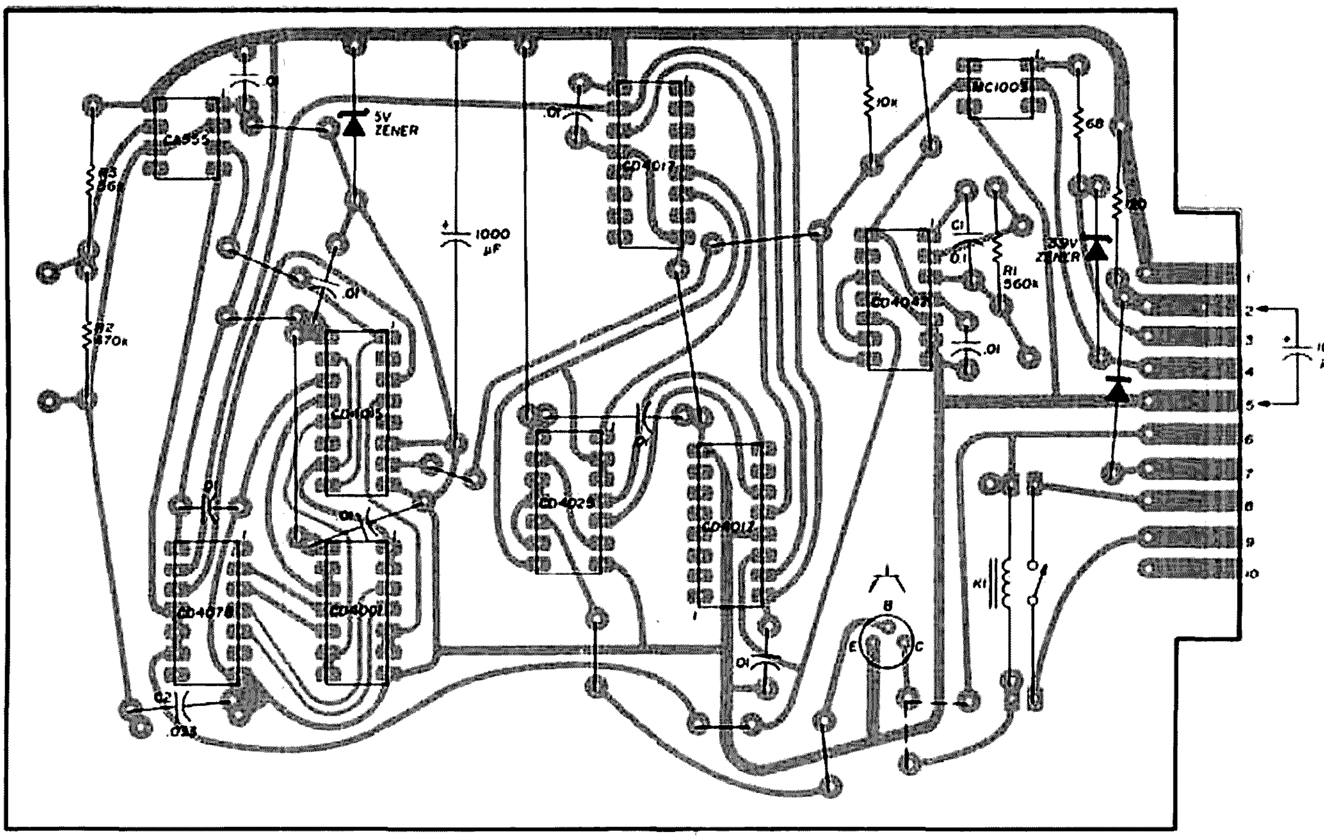


fig. 2. Printed-circuit layout for the RTTY line-end indicator. The unused contacts at R1, R2 and C2 are used, as required, for frequency adjustment. The connection from the collector of the transistor Q1 is chosen for relay or lamp option (see fig. 1). Connect the lamp to pins 2 and 6 of the board. For external power supply, omit the external 100 µF capacitor, 120 ohm resistor and 5 volt zener and short pin 1 of the board to pin 2.

Now that the lamp has gone on, a few more letters may be typed and then *carriage return-line feed-letters*. The *carriage return* code group is used to reset gate only when the *carriage return* character (SSSMS) is struck. The RCA CD4015 is a serial input, parallel output, static shift register. Each time a

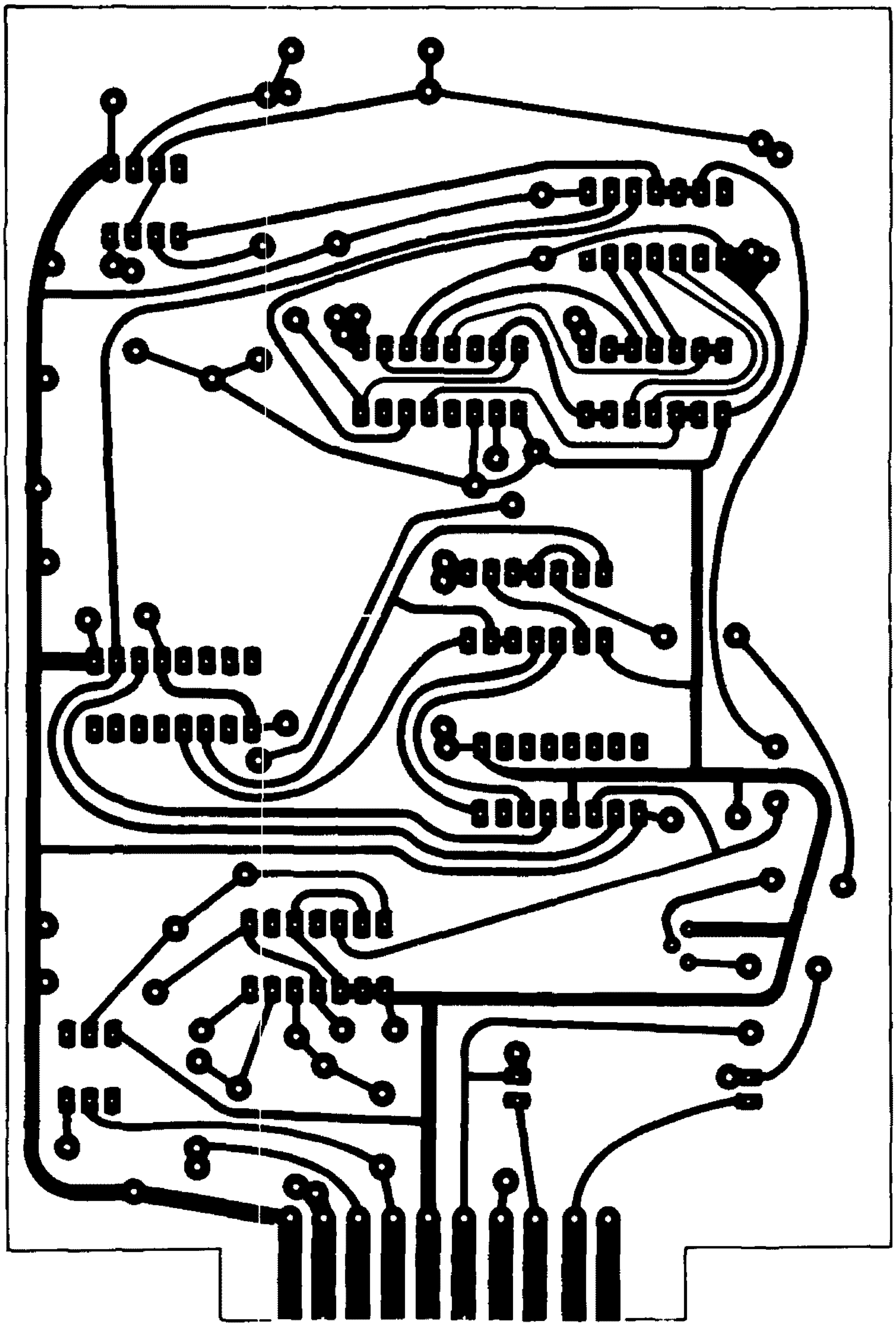


fig. 3. Full-sized printed-circuit board for the RTTY line-end indicator. Component layout is shown in fig. 2.

the counters to zero for the next line. Here's how it works: A shift register, U5, used to read the five-unit Baudot code, is preset to trigger another NOR pulse appears at its clock input, the information already in the register is shifted over one stage. The parallel output allows data monitoring at each stage.

The clock pulses are provided by an RCA CA555 astable timer, U9, set for approximately the same pulse width as the 22 *millisecond Baudot code. The* start pulse at the output of the optical isolator will turn on the monostable oscillator, U2, and will also appear at the data input of the shift register, U5. The oscillator output, in turn, will trigger the astable timer, U9, but will allow it to stay on only for seven pulses. These seven pulses will clock the register for each of the seven mark, or space, pulses of a Baudot character. After seven pulses the register will contain the code for the character sent.

Ignoring the start and stop pulses, the five-character code pulses available at the parallel outputs can now be checked to see if they form the code group for *carriage return* (SSSMS). By placing inverters in each of the space outputs, all outputs applied to U8 (RCA CD4078) become low. This condition will occur only for *carriage return* (for any other character, one or more of the outputs will be high). With all inputs low, the output of U8 will go high and reset U3 and U4. While the seven pulses are being fed into the shift register, the output of U8 is kept low by the input lines (pins 9, 10, 11) which are tied to the high output of U2.

If you were concerned about the lamp going on too close to the line-end after character 68 was struck, relax. All amateurs send the *carriage return-line feed-letters* combination at the end of line. The last two functions do not move the type box, so the lamp will go on at 66 printed characters, the same as the mechanical switch on the printer. Use of the *figures* or *letters* key on any line of type provides a safety factor. They will be counted even though there is no print.

Construction is simple whether you use hand wiring or a printed-circuit board. The layout of the circuit board is shown in **fig. 2**. The board can be mounted inside the machine. The printer remains in the TU loop while the keyboard and line-end indicator input are put in series in a separate loop to the perforator (watch the polarity).

If the board is too big, as shown, you can build a smaller one as there really is nothing critical. The only adjustments are to the two oscillators. The values shown in **fig. 1** should work with no problem. For exact adjustment use a digital counter and vary R1 and/or R2 for the monostable oscillator, U2, and astable oscillator, U9, respectively. Set the monostable for 150 milliseconds (7 Hz) and the astable for 22 milliseconds (45 Hz). If in doubt, set the astable slightly on the low frequency side. The rise time of its pulses can occur anytime during each 22 millisecond Baudot pulse.

A relay to turn on an existing margin light may be substituted for the lamp shown in **fig. 1**. The PC board provides contacts for a relay such as the General Reed GR410-P5 or Clare MRB 1A05. Just wire the relay contacts in parallel with the machine margin switch.

The circuit board also provides for a rectifier diode to allow use of 6.3 volt ac power. A dropping resistor and 5.1 volt zener diode are added to prevent wide voltage swings as the lamp turns on and off.

conclusion

And what do you have after you built this CMOS line-end indicator? You can receive a message and while reading what is being printed, you can type answers to questions, ask questions, or make comments. When the other station signs, you can turn on the transmitter and TD and continue to punch tape while you are transmitting commercial quality, 60 wpm copy.

reference

1. H. Dressel, W2UVF, "RTTY Line Length Indicator," *ham radio*, November, 1973, page 62.

ham radio

tunable audio filter

for weak-signal communications

A discussion of weak-signal CW detection techniques, including a versatile, high-performance audio filter

Ken Holladay, K6HCP, Gilroy, California 95020

Although there has recently been a great deal of controversy over the value of CW communications, CW continues to be the most efficient mode of radio transmission. This was true in the first successful transatlantic tests of the 1920s and it is still true today as amateurs work to improve the reliability of EME, meteor scatter and long-range tropospheric communications. Once a communications path has been proven to exist with CW, amateurs usually find a way to use ssb (or other mode) over the same path. CW is also valuable for working DX on the high-frequency bands when propagation conditions are poor, particularly during periods of low solar activity.

Amateur communications equipment has come a long way since 1921 when Paul Godley set up a receiving station

on the Scottish coast to listen for signals from the United States. Now our transmitters run higher power, are more efficient and are stable; high-gain, directional antennas are commonplace and our stable and sensitive communications receivers are equipped with highly selective mechanical or crystal filters.

The communications receiver conditions the incoming CW signal so that the greatest receiver of all – our ear-brain – can capture that first detectable signal. The ear is capable of receiving signals from less than 20 Hz to more than 20 kHz and, when coupled with the brain, forms an extremely efficient and versatile CW detector. Furthermore, the ear-brain combination is capable of acting like a variable-frequency, variable-bandwidth audio filter which allows us to detect and copy signals which are buried in the noise or nearly obliterated by interference.

The human ear is actually able to hear signals which are *below* the noise level.[1] Tests conducted by W2IMU, using a 3 kHz bandwidth receiver and a signal generator, showed that when a CW signal is adjusted to the same audio level as the noise (zero dB signal-to-noise ratio), the signal was 100 per cent

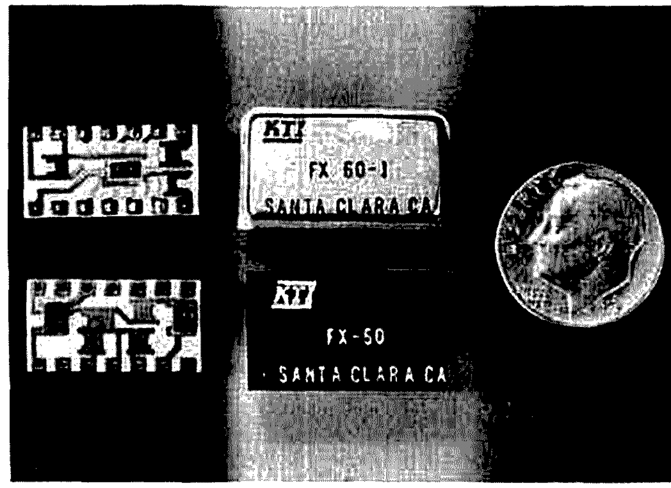


fig. 2. Universal Active Filters manufactured by KTI.

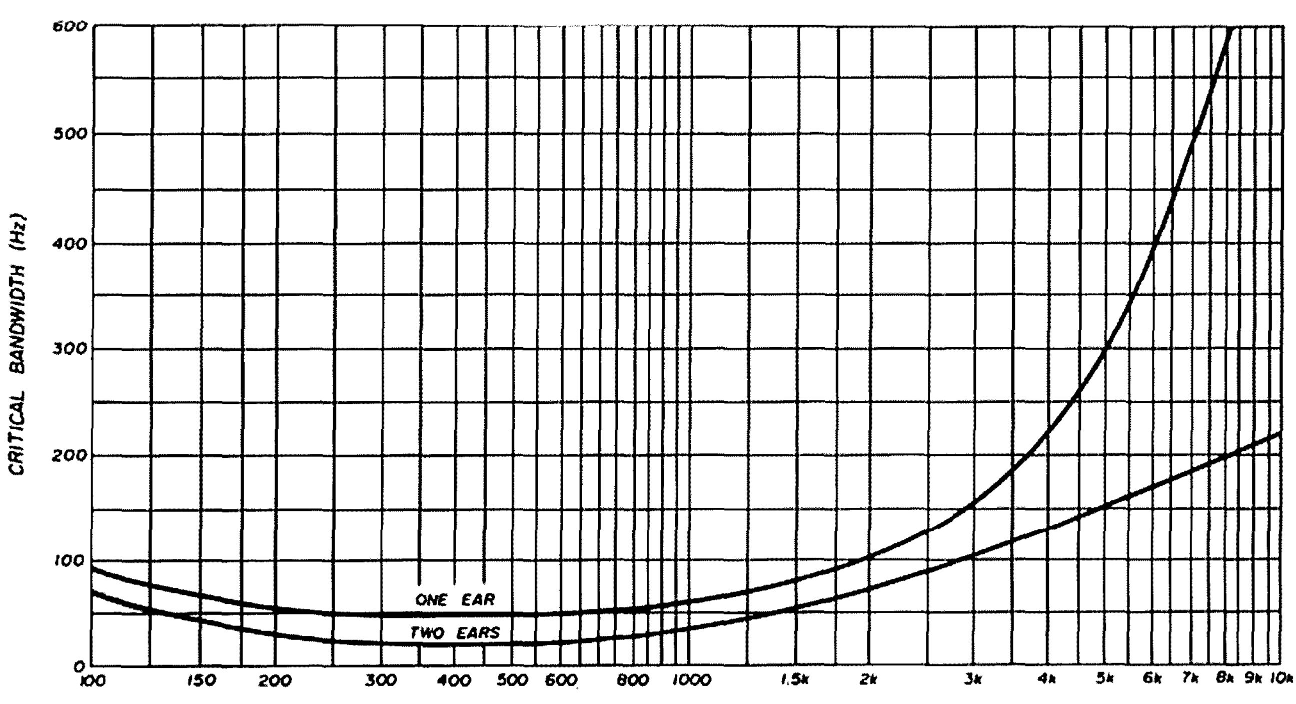


fig. 1. Critical bandwidth of the human ear as a function of frequency.

readable. The input signal was then reduced in 3 dB steps. Copy became more difficult but callsigns could still be accurately identified at 9 to 12 dB *below* the noise level. Although the *presence* of signals 20 dB below the noise could still be detected, the signals could not be copied.

The reason that these weak signals can be copied reliably is that the ear-brain filter has narrowed its bandwidth to approximately 50 Hz! The graph of **fig. 1** *shows the frequency response of* the human ear vs its bandwidth.[2] This curve also shows that 1000 Hz is *not* the optimum tone with which to copy weak CW signals. Most amateurs who have worked with weak CW signals have found that they prefer a lower pitch as signals get weaker. **Fig. 1** shows why.

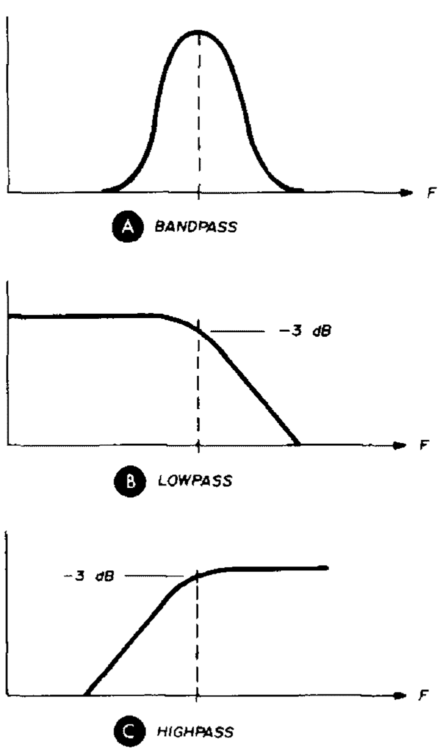


fig. 3. Comparison of active filter outputs as a function of frequency.

Another reason to lower the frequency of the signal you want to copy is that, if there is interference, the lower-frequency signal is easier to detect. For example, if the frequency difference between the desired and undesired signals is 100 Hz, and the desired signal is tuned for a 1000 Hz pitch, the frequency difference is only 10 per

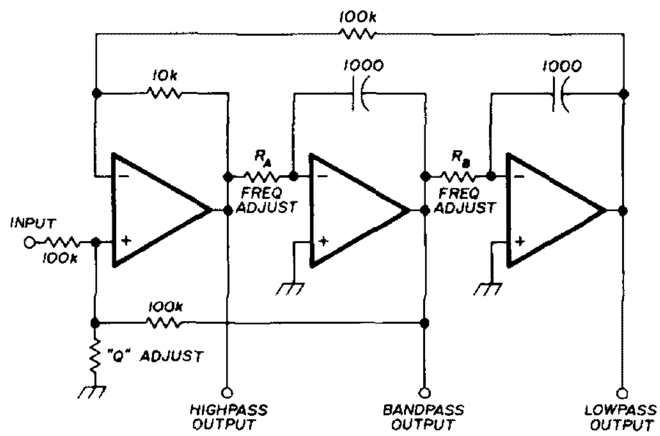


fig. 4. Basic schematic of the KTI Universal Active Filter.

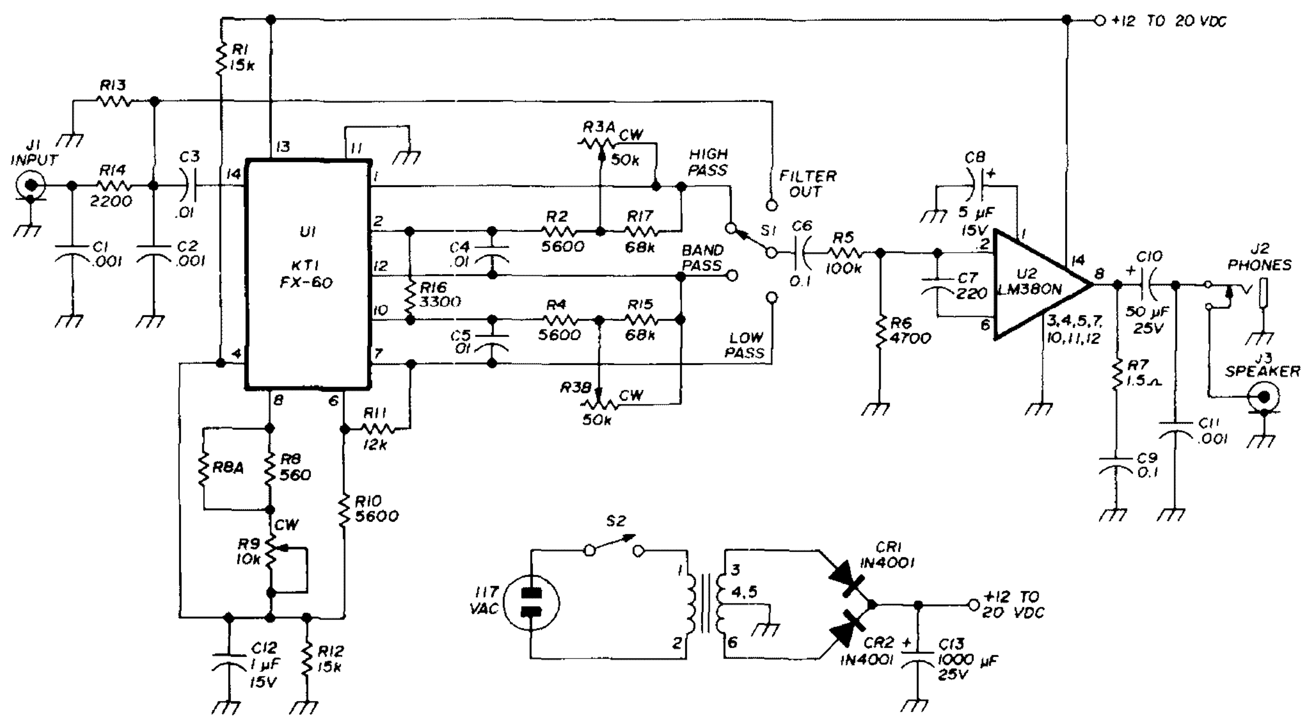
cent. If you tune the desired signal for a 500 Hz pitch, the frequency difference is increased to 20 per cent, a 2:1 improvement.

The human ear-brain also copies signals by comparing signal against signal or signal against noise. If a narrow bandpass filter, say 200 Hz, is used in the receiver it excludes other signals as well as some of the noise. This is fine for strong signals but causes problems with weak ones because too much bandwidth restriction limits the amount of noise the ear has to compare with.

Very sharp filters also have a tendency to "ring" – this ringing sounds much like the signal and makes signal-to-noise comparison difficult, if not impossible, with very weak CW signals. In addition, narrow bandwidth filters are usually tuned to some fixed frequency so the individual operator cannot optimize the frequency and bandwidth of the filter to complement his own ear.

Since the human ear is already capable of 50 Hz bandwidth, very narrow filters are not the best for weak CW detection except for eliminating inter-

ference. What is needed is a variable frequency and variable bandwidth filter that can be adjusted for various operating conditions. Variable audio filters are sponse to less than 10 Hz. They can also provide simultaneous highpass, bandpass and lowpass outputs as shown in **fig. 3** so they are ideal for such applications as



R3	Frequency adjust. Dual 50k potentiometer, CCW log taper (Allen-Bradley 70C1N048-503B or equivalent)
R9	Bandwidth adjust. 10k potentiometer, linear taper
R8A	Selected for highest Q (narrowest bandwidth)
R13	Sets input level (10V p-p maximum), not normally required
S1	4 position, non-shorting rotary switch
T1	24 volt, 180 mA power transformer (Signal PC 24-180)

fig. 5. Tunable audio filter uses KTI FX-60 Universal Active Filter and provides highpass, bandpass and lowpass outputs. Circuit has unity voltage gain so it may be switched in and out of the receiving system as required without adjusting the audio gain control. Printed-circuit layout for the filter is shown in fig. 6.

difficult to build with lumped values of inductance and capacitance, but modern integrated-circuit technology provides the basis for excellent audio CW filters. Kinetic Technology* has developed a line of Universal Active Filters which can be used from less than 1 Hz to greater than 100 kHz, depending on the model. The bandwidth of these active filters can be adjusted for a flat response to less than 10 Hz.

*KTI/Division Baldwin Electronics, Inc., 3393 De La Cruz Boulevard, Santa Clara, California 95050, telephone (408) 296-9305.

speech filters, notch filters, tone encoders, RTTY and, best of all, CW filters.

tunable cw filter

The KTI active filters use three operational amplifiers in a stable, negative-feedback circuit (**fig. 4**) which is commonly called a bi-quad. Although a complete description of device operation and its various connections is beyond the scope of this article, complete data is available from KTI.

A tunable CW filter which uses the KTI FX-60 and an LM380 audio ampli-

fier is shown in fig. 5. This filter tunes the audio range from 300 to 1800 Hz and its bandwidth can be adjusted from 50 to 1200 Hz. The filter, which has unity gain and is built on a printed-circuit board, is designed to be plugged into the headphone jack of a communi-

(C1, R14, C2) passes the audio frequencies but blocks rf energy. Resistor R13 is used to lower the input signal level, if required, to the FX-60 active filter. The dual 50k potentiometer, R3A and R3B, sets the frequency of the filter while the bandwidth is adjusted with

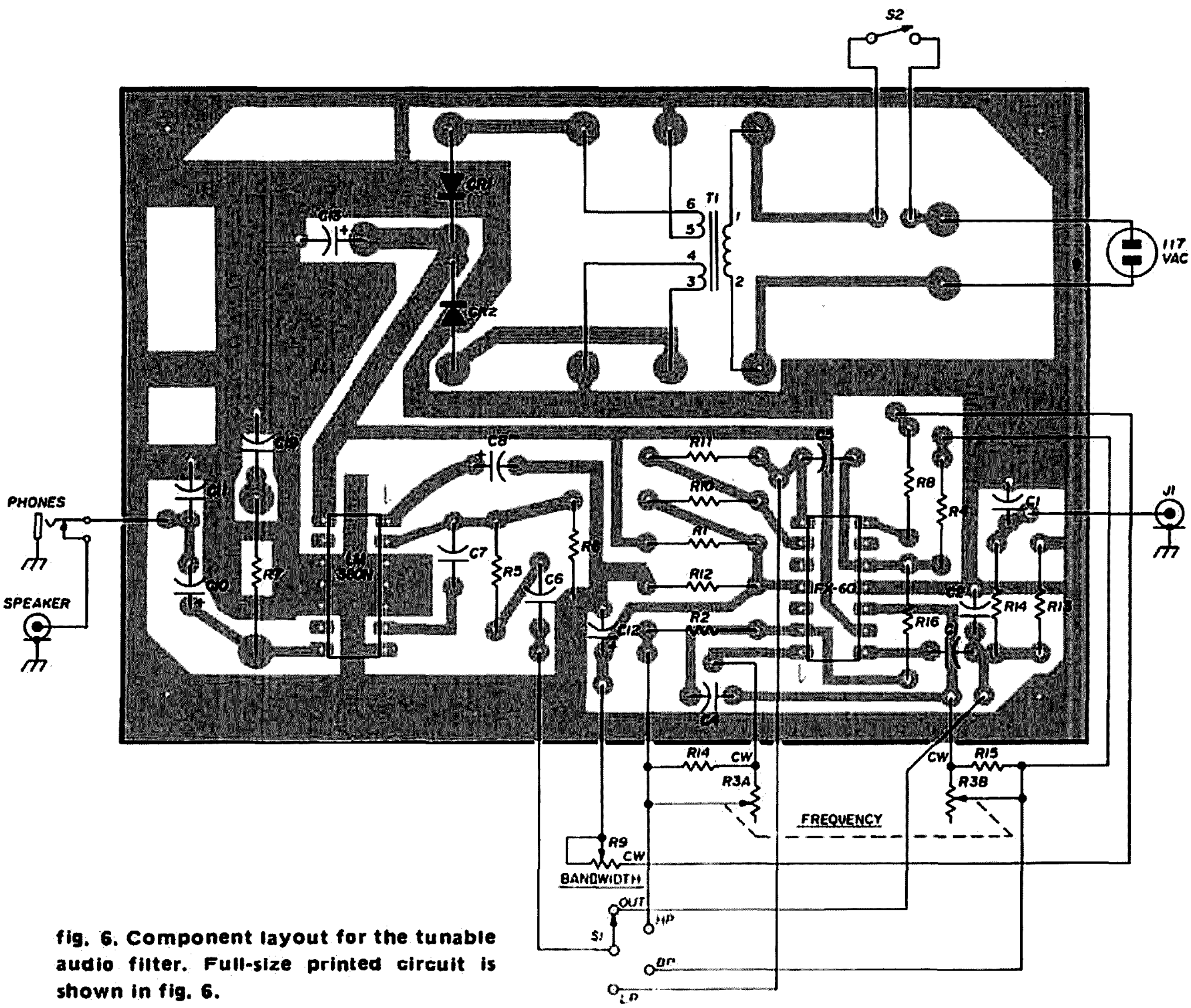


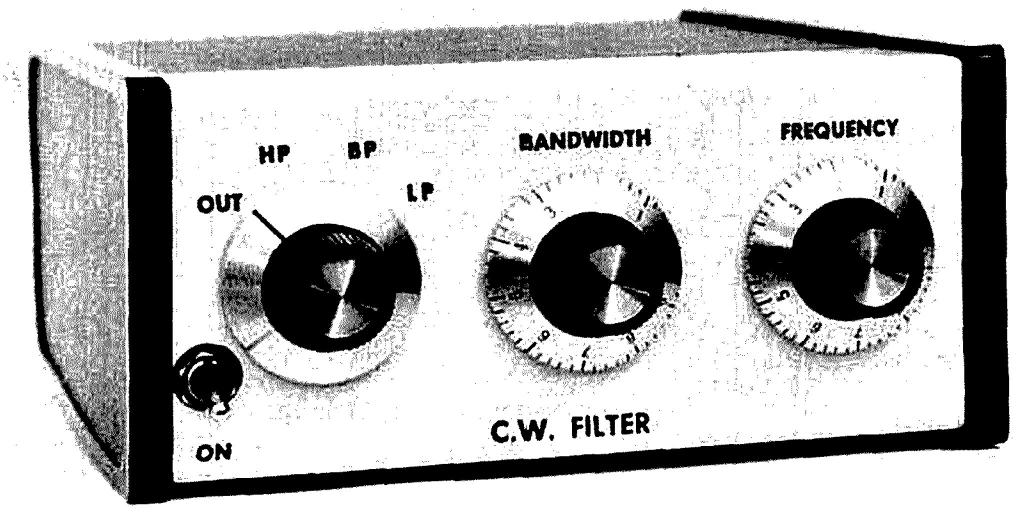
fig. 6. Component layout for the tunable audio filter. Full-size printed circuit is shown in fig. 6.

cations receiver. The output is then connected to the speaker or headphones and the filter can be switched into the circuit as required. The LM380N provides two watts of audio output, more than enough for most applications.

In the circuit of fig. 5 the audio signal from the receiver is introduced to the CW filter at J1. The input pi network

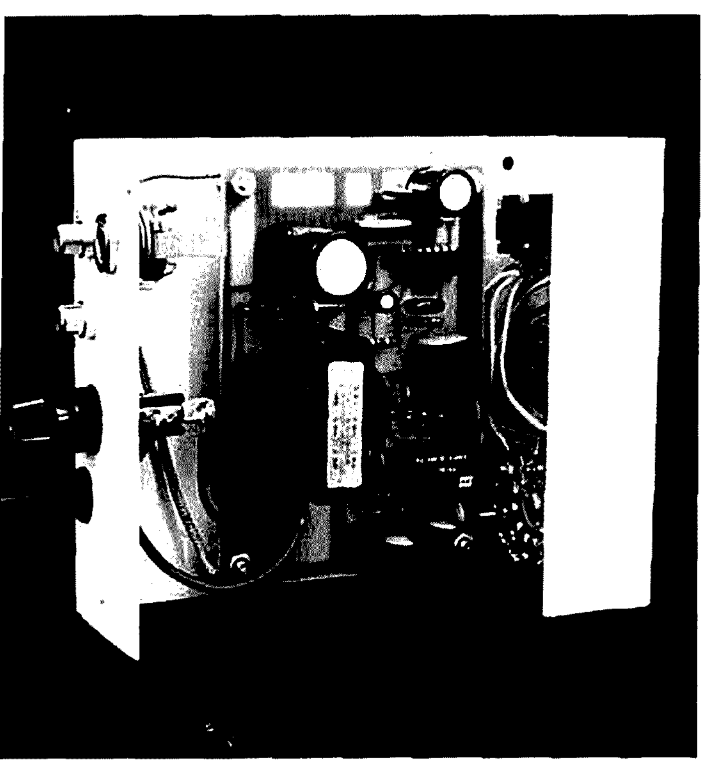
potentiometer R9. The function switch, S1, selects the highpass, bandpass or lowpass output from the FX-60 or switches the filter out of the circuit.

In the active filter circuit resistors R1 and R12 provide the necessary biasing so the FX-60 can be operated from a single, positive power supply. Resistor R11 allows the three outputs to be at



Tunable audio filter built by W1DTY is housed in Ten-Tec JW-5 enclosure. Input and output jacks are on rear panel.

Construction of the tunable audio filter built by W1DTY using printed-circuit board available from Holladay Communications. Input and output jacks are on rear panel, left. Board is installed on chassis with 0.25" (7mm) spacers.



the same level. R10 limits the widest bandwidth while R9 sets the narrowest limit.

During setup resistor R8A is adjusted until the circuit goes into oscillation; the correct value is that just before the circuit oscillates. The narrowest bandwidth will vary from unit to unit, and some may not require R8A. Resistor R16 maintains filter stability at the narrow bandwidth setting and capacitors C4 and C5 set the frequency range.

The National LM380N audio power IC is connected to the function switch through the dc blocking capacitor, C6. Resistors R5 and R6 set the input level and capacitor C7 provides high-frequency rolloff at 4 kHz. The series RC circuit (R7, C9) from the output pin to ground prevents high-frequency oscillations.

The tunable audio filter is built on a 3 by 4.4 inch (7.6 by 11.2cm) printed-circuit board. The component layout is shown in **fig.** 6. Printed-circuit boards

and special components are being made available in conjunction with this article.*

The tunable audio filter may be used to improve various types of receiver signals. In the lowpass mode it can be helpful with ssb reception. For use on CW it should be set to the *bandpass* position, adjusted to narrow bandwidth and peaked on the desired CW signal. The optimum frequency and bandwidth will vary from operator to operator, as discussed previously. Operation of the unit is quite simple and there is no preset adjustment to follow. Some amateurs like to use the unit in the narrow bandwidth, lowpass mode (**fig. 8**) as this provides some low-frequency noise to which the ear can compare weak CW signals.

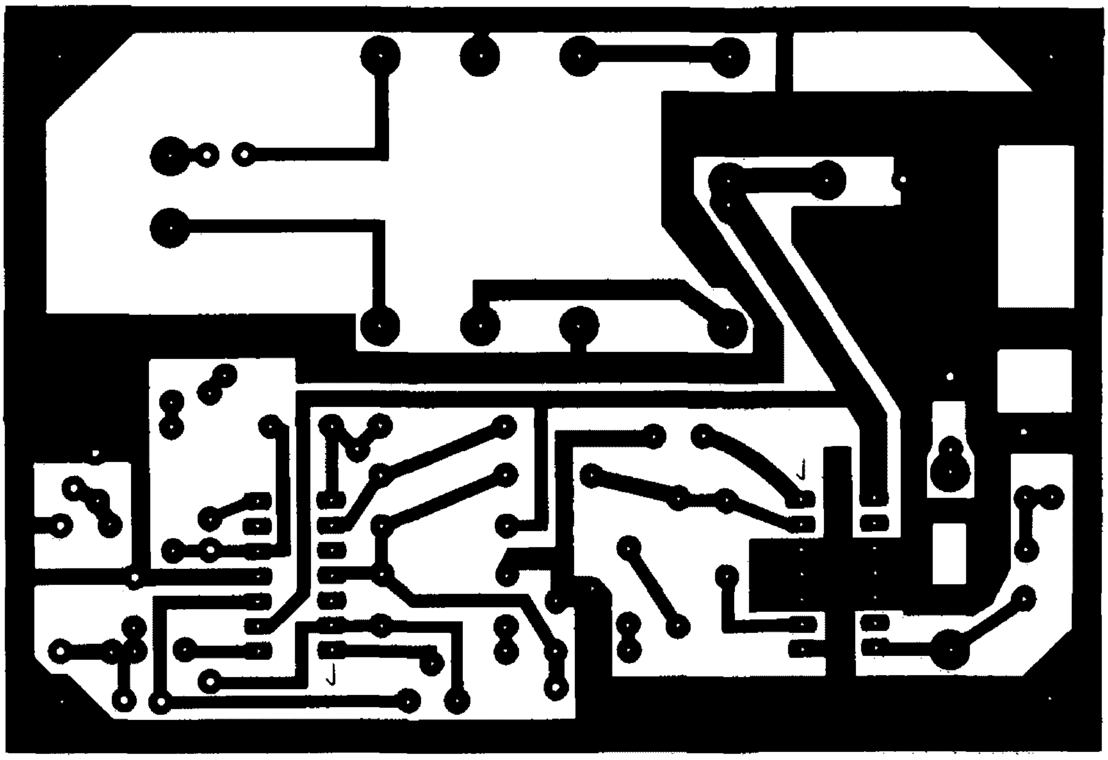


fig. 7. Full-size printed-circuit layout for the tunable audio filter. Drilled PC boards are available (see footnote below).

fig. 8. Frequency response of the tunable audio filter set for narrow bandwidth in the lowpass position. This response is sometimes preferred for weak-signal CW work.

*The following components can be supplied: drilled and plated printed-circuit board, $5.75; KIT FX-60 Universal Active Filter, $6.95; National LM380N audio power IC, $1.75; Allen-Bradley dual 50k potentiometer, CCW log taper, $7.30; power transformer, Signal PC 24-180, $4.80. Wired and tested filters, model AF-100, complete with enclosure are also available for $60.00. Order from Holladay Communications, 2140 Jeanie Lane, Gilroy, California 95020.

references

1. R. Turrin, W2IMU, "Simple Super Selectivity," *QST*, January, 1967, page 48.
2. P. Laakmann, WB6ION, "Signal Detection and Communication in the Presence of White Noise," *ham radio*, February, 1969, page 16.

ham radio

frequency selective and sensitivity controlled sstv preamplifier

Discussion of a specially designed op-amp circuit for reception of sstv pictures under adverse operating conditions

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Generally the **input circuit** of a sstv receiver uses a limiting amplifier so the frequency-modulated sstv signal is independent of amplitude variations. If this amplifier is not frequency selective and not sensitivity controlled, all signals within the range of the sstv signal frequency with amplitudes high enough to be limited will pass through the circuit and influence the picture.

Some sstv receivers use a filter in the input stage, but its efficiency depends on the signal strength, and undesired weaker signals, including unwanted sstv signals at the same frequency, will not be suppressed. A frequency-selective and sensitivity-controlled limiting amplifier avoids these disadvantages. It can be used as a preamplifier with any sstv receiver and permits the reception of sstv pictures under extremely adverse conditions.

In the block diagram of the system shown in fig. 1, the linearly amplified input signal passes through a high-pass filter and appears as a square-wave signal at the output of the comparator *only* if its amplitude exceeds that of the reference voltage, which is proportional to the peak voltage of the 1200-Hz synchronizing signal.

With this circuit the cutoff frequency of the high-pass filter is independent of the input amplitude. Furthermore, the sensitivity of the comparator is adapted to the amplitude of the sstv signal. Weaker signals, including sstv signals which are weaker than the desired signal, are totally suppressed.

Fig. 2 shows the circuit in detail. The back-to-back diodes (CR1 and CR2) at the input protect the linear amplifier, U1, from excessive drive. The high-pass filter (U2 and U3) is an active Tschebyscheff filter of the order n = 4 with a cutoff frequency of about 1000 Hz (-60 dB/decade).

The frequency of the selective amplifier, U4, can be tuned to 1200 Hz by means of potentiometer R1. The peak voltage detector (U5 and U6) has a charging time constant of about 1 millisecond; the discharging time constant was chosen to be about 1 second.

The comparator, U7, is clamped by two back-to-back diodes to limit the output amplitude to 0.7 volt. In most cases this amplitude is still too high and

*A printed-circuit board and parts kit are available from the author.

must be further reduced, typically to 100 millivolts or so. The output can be adjusted to the required level with potentiometer R4.*

To adjust the system first short the input of U1 and set the output of U6 to zero voltage with potentiometer R2. Feed an sstv signal into the input and adjust R1 for maximum reference voltage. Now reduce the reference voltage with R3 as much as is required to synchronize the picture. Otherwise the

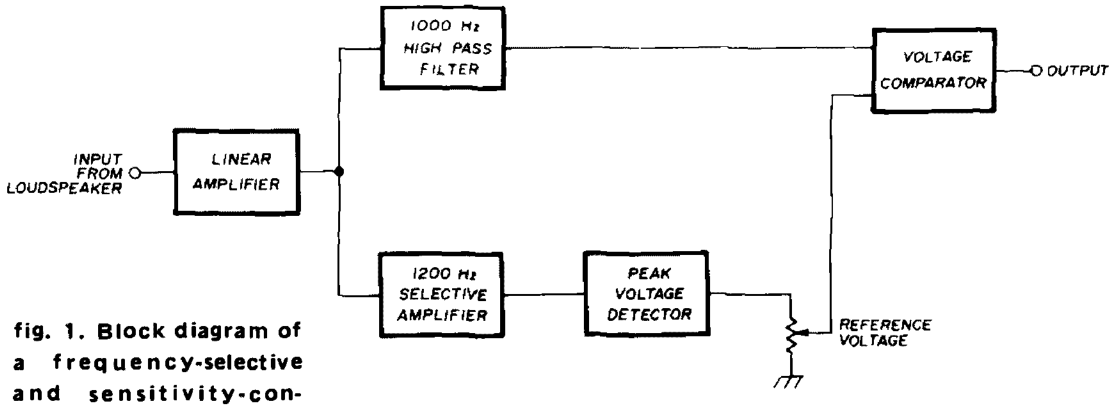


fig. 1. Block diagram of a frequency-selective and sensitivity-controlled sstv preamplifier which permits reception of sstv signals under adverse conditions.

table 1. Operating specifications for the frequency-selective and sensitivity-controlled sstv preamplifier.

Input sensitivity	1 mV rms
Regulating range	1 mV to 500 mV rms
Frequency range	1000 to 2800 Hz
Power requirements	±15 volts, 16 mA

reference voltage is too high and the synchronizing frequency cannot pass through the comparator. Complete specifications for this circuit are given in **table 1.**

ham radio

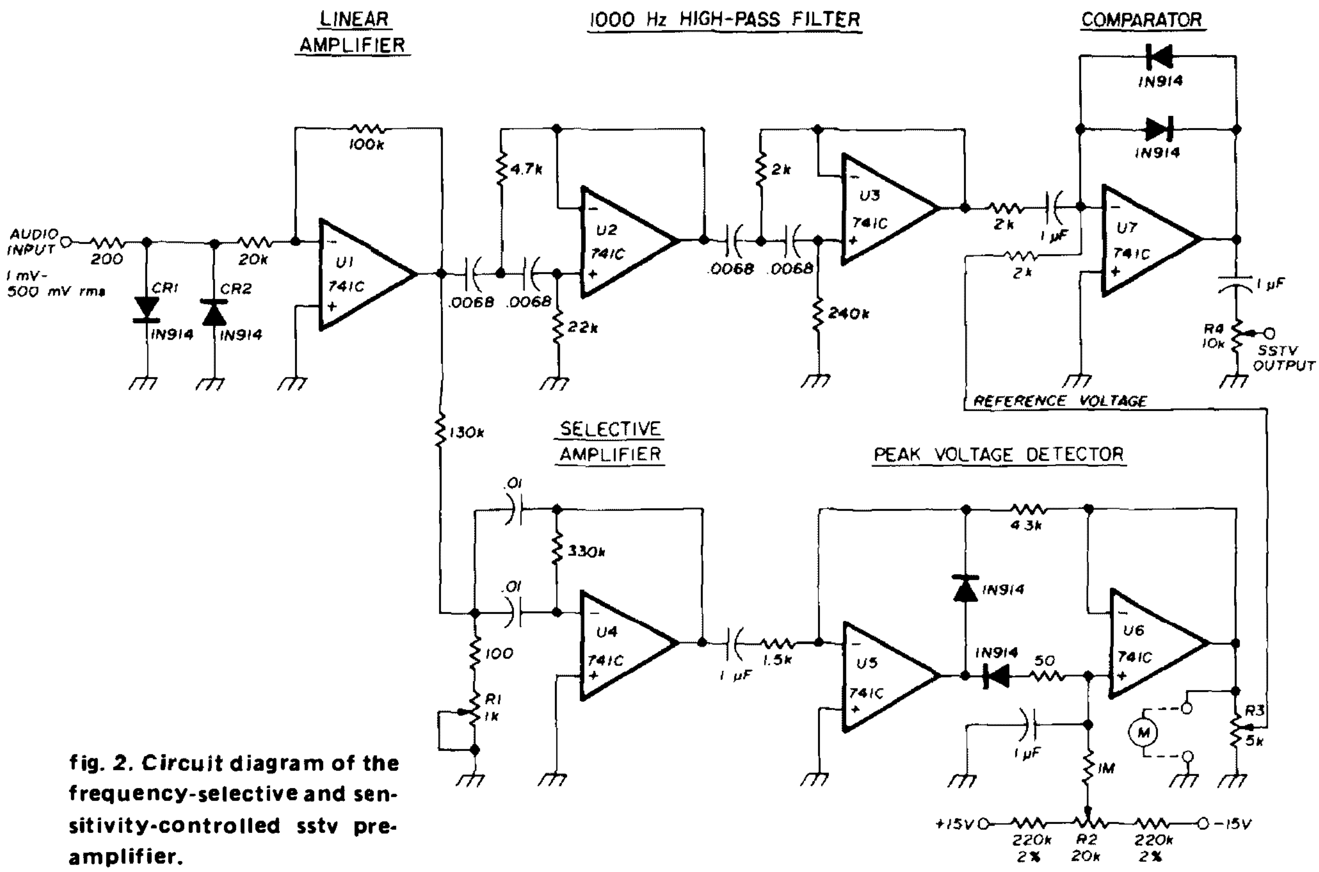


fig. 2. Circuit diagram of the frequency-selective and sensitivity-controlled sstv preamplifier.

the crystal mixer:

recipe for curing receiver drift

A novel method for taming receiver local oscillators, featuring a VXO and a divide-by-*n* network

Bill King, W2LTJ, 5 Midwood Drive, Florham Park, New Jersey 07932

RTTY operation made me aware of the need for frequency stability in my receiver. When copying nets receiver drift, although small, became a big problem. For example, autostart nets demand receiver stability on the order of several Hz. I tried the usual remedies to improve receiver frequency stability: voltage regulation, ventilation, and reduced heat sources but more stability was needed. Crystal control was the obvious answer.

the rock-mixer

My approach to the problem was simple and inexpensive. The device described here, which I call the rock-mixer, uses only a handfull of ICs and very few crystals. In fact, one odd-ball crystal and the rock-mixer will allow you to tune the entire 20-meter band with crystal control all the way. The rock-mixer is easy to adapt to almost any receiver — no phase-locked loops, no filters, and no other synthesizer-type complications.

The rock-mixer creates two frequencies and combines them to produce a pulse train with which your receiver local oscillator can synchronize. Your job is to determine what combination of frequencies to use, given the modest selection of crystals you might have on hand. You don't even need crystals in

table 1. N values and lock-on frequencies produced when the receiver local-oscillator frequency is above the crystal frequency. F_x = 15000 kHz, $F_{i\text{-}f}$ = 1650 kHz.

n	divide-by-n output, F_x/n (kHz)	LO frequency $[F_x + (F_x/n)]$ (kHz)	received frequency $[F_x + (F_x/n)] - F_{i\text{-}f}$ (kHz)
19	789.474	15789.474	14139.474
20	750.000	15750.000	14100.000
21	714.286	15714.286	14064.286
22	681.818	15681.818	14031.818
23	652.174	15652.174	14002.174

F_x = crystal frequency 15000 kHz
$F_{i\text{-}f}$ = intermediate frequency = 1650 kHz

the amateur bands; surplus crystals work fine and are inexpensive.

Fig. 1 is a version of the rock-mixer which includes a variable frequency crystal oscillator, a divide-by-*n* circuit (VXO), a mixer NAND to combine the fundamental and divided frequencies, and a coupling capacitor to your receiver local oscillator (LO). The coupling capacitor can be a gimmick (wire twisted around the grid lead to the LO tube).

operation

The VXO provides a stable frequency, which is tunable over a modest range. The output of the divide-by-*n* is added to or subtracted from the VXO frequency to produce a pulse train from the mixer NAND to which the receiver LO can lock. Example: suppose you wish to tune in a station at 14062 kHz and the crystal you choose is 15000 kHz. To tune 14062 kHz with an i-f of 1650 kHz requires 15712 kHz at the receiver local oscillator, which is a frequency difference of 712 kHz. To get a ballpark value for *n*, divide 15000 by 712, which equals 21.067. Note, however, that only whole numbers can be used in the divide-by-*n* circuit, which must provide a pulse train with a frequency that the receiver local oscillator can lock onto to produce the frequency to be received (14062 kHz).

Table 1 shows what can be expected for *n* between 19 and 23, for example, in the 20-meter band when the local oscillator frequency is above that of the VXO crystal. Note that the integer 21 yields a received frequency of 14064 kHz. But since we wish to receive 14062 kHz, we must find a frequency to which the 15000-kHz crystal can be VXOed to produce 14062 kHz.

When the local oscillator frequency is above the crystal frequency,

$$F_{\Delta x} = \frac{(F_d + F_{i\text{-}f})\,n}{n+1} \qquad (1)$$

where

$F_{\Delta x}$ = crystal frequency changed by the VXO

F_d = desired frequency (14062 kHz)

$F_{i\text{-}f}$ = intermediate frequency (1650 kHz)

n = number to which the divide-by-*n* circuit is set (21 in the example)

If the local oscillator frequency is below the crystal frequency, simply replace $n + 1$ in the denominator of **eq. 1** with $n - 1$.

Using the example above, in which the local oscillator frequency is above

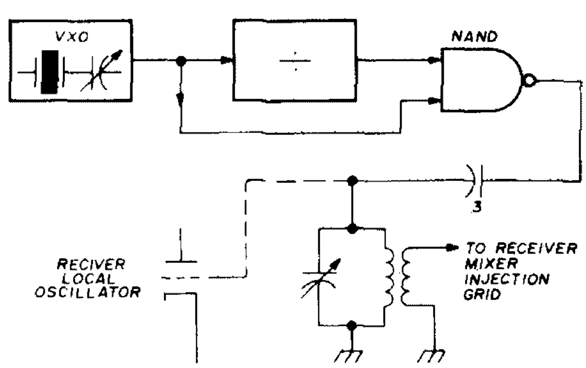


fig. 1. Rock-mixer block diagram. Circuit features surplus crystals and readily available ICs. The value is determined by formula to produce a synchronizing pulse train for the receiver LO.

the crystal frequency, then the frequency to which the crystal must be VXOed to receive *14062 kHz is* 14997.818 kHz. By VXOing the 15000-kHz crystal to 14997.818 kHz and setting the divide-by-*n* circuit to 21, a pulse train of 15712 kHz will be produced from the mixer NAND to which the receiver local oscillator can lock onto to bring in the station at 14062 kHz.

table 2. Receiver frequency ranges for the *20-meter* band as a function of n, using a 15-MHz crystal.

n	frequency tuned by each n (kHz)
21	14001 to 14077
20	14037 to 14113
19	14076 to 14152
18	14120 to 14196
17	14169 to 14245
16	14224 to 14300
15	14286 to 14363
14	14357 to 14434

You'll note that other frequencies will appear in the mixture feeding the local oscillator, but the predominant frequency is the crystal frequency plus the divide-by-*n* output (see **table 1**).

Now let's review what we did. The 15-MHz crystal frequency has been changed by the VXO to 14997.818 kHz. Since *n* is 21, the divide-by-n output is 714.182 kHz; therefore, the sum *and difference frequencies are* 15712.000 and 14283.637 kHz, respectively.

To attain lock is a simple matter. Tune the dial to a point near the expected place. As you approach it the receiver will lock from as far away as 50 kHz under the right conditions. When a lock has been attained you can tune the receiver knob a modest amount on either side and not lose the desired station. *When you exceed the lock-in* range you will hear squishing noises as well as other stations, but you won't lose the desired station. If you desire to fine-tune the station, then tweak the VXO knob and the receiver will follow.

When the lock-in condition exists, the receiver tuning will be under complete control of the VXO: thus the VXO knob is now the receive tune knob. The tuning range will be limited to the extent that you can "rubber" the crystal. For a 15-MHz crystal my VXO covers 14940 to 15012 kHz. **Table 2** shows the corresponding ranges that can be covered on 20 meters with the 15-MHz crystal in the VXO and with the divide-by-*n* circuit set as shown. (Remember the i-f was 1650 kHz.)

table 3. Selected frequencies for RTTY using the SX-100 receiver (1650-kHz i-f).

frequency (kHz)	station	speed/shift (wpm/Hz)	n	crystal frequency (MHz)
3600	Autostart net	60/850	22	5.02
3623	W1AW ARRL Headquarters	60/170-850	20	5.02
3223	WBR70 Miami WX	60/850	12	9.00
8105	WBR70 Miami WX	60/850	12	9.00
8105	WBR70 Miami WX	60/850	14	9.10
8105	WBR70 Miami WX	60/850	40	10.00
12175	WBR70 Miami WX	60/850	13	15.00
12175	WBR70 Miami WX	60/850	140	13.92
8183	UPI News in English	66/550	11 or 22	9.01
19537	AP News in English (NYC)	66/400	6 or 12	9.08
5460	Voice of America (USIS)	60/400	64	7.00
5460	Voice of America	60/400	9	8.00
10972	Voice of America	60/400	12	11.65

Table 2 shows that the entire 20-meter band can be received, crystal controlled, using only one crystal. In my case a 15-MHz rock was used, but note that almost any crystal near 15 MHz will work. For example, a 15239 kHz crystal or another oddball like 14875 kHz would work just as well; the only change would be the *n* value required.

You don't need to make a fundamental pulse train, because harmonics and subharmonics are almost as good. In the examples with two *n* values the oscillator is made to lock onto frequencies twice removed from the usual frequencies. Consider a 10-MHz VXO and an *n* value of 10. The receiver oscillator can just as easily lock onto 9.0, 10.0,

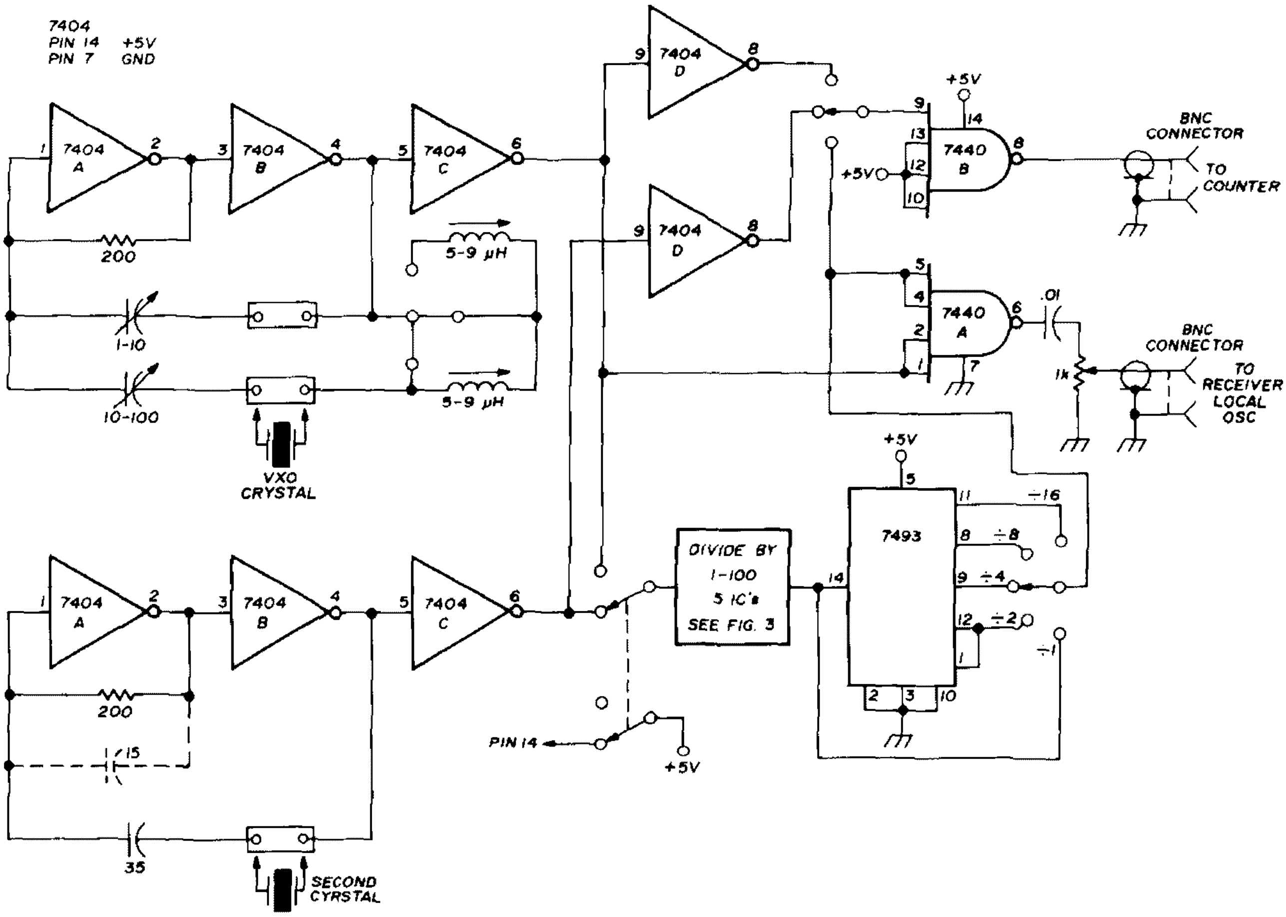


fig. 2. Circuit for one or two crystals. A pair of 7404s form the VXO and the second crystal oscillator. A 7440 drives a counter and provides mixing for the divide-by-n circuit.

Table 3 shows several frequencies I use for RTTY. These are real examples, so included in the list are the *n* values and the nominal crystal frequency to tune in the particular station. Some examples in **table 3** are straightforward and some are a little tricky. The one at 3223 kHz is a case in point. The 9-MHz crystal and the divide-by-12 circuit produce 9746 kHz to which the receiver LO at 4873 kHz can easily lock (i-f is 1650 kHz).

and 11.0 MHz. Now make *n* = 20 and lock-on to the same frequencies occurs as well as to many others, such as 9.5, 10.5, 11.5 MHz, etc. There are other reasons to use the doubled *n* value, which are treated later.

two-crystal rock-mixer

There may come a point when you give up trying to find a good combination of *n* with a crystal you have available. Enter the two-crystal rock-mixer.

In this arrangement you have one crystal in the VXO and a different crystal feeding the divide-by-*n* circuit; thus, you have greatly expanded the rock-mixer capability.

Fig. 2 shows the rock-mixer circuit in which one or two crystals can be used. Two 7404 hex inverters form the VXO[1] and the second crystal oscillator. Two variable capacitors are shown with two crystal sockets. It is important to keep circuit stray capacitance at a minimum when varying the crystal frequency in the high direction. Note that no switch is used for changing crystals, because it's not possible to keep stray capacitance to a low enough value with any type of switch.

The divide-by-*n* circuit consists of two parts: a set of two 74195 shift registers (**fig. 3**) and a 4-bit binary IC (7493) connected to divide by 1, 2, 4, 8, 16. The 7493 is a "division range increaser," but more important is its function as a "duty-cycle improver." **Table 4** shows the application of the 7493 for this purpose.

Fig. 3 depicts the shift registers and the logic to divide by any number from 1 to 100. This circuit was found in the Fairchild *TTL Applications Handbook*.[2] The only changes I made were to substitute some equivalent ICs instead of those in the reference.

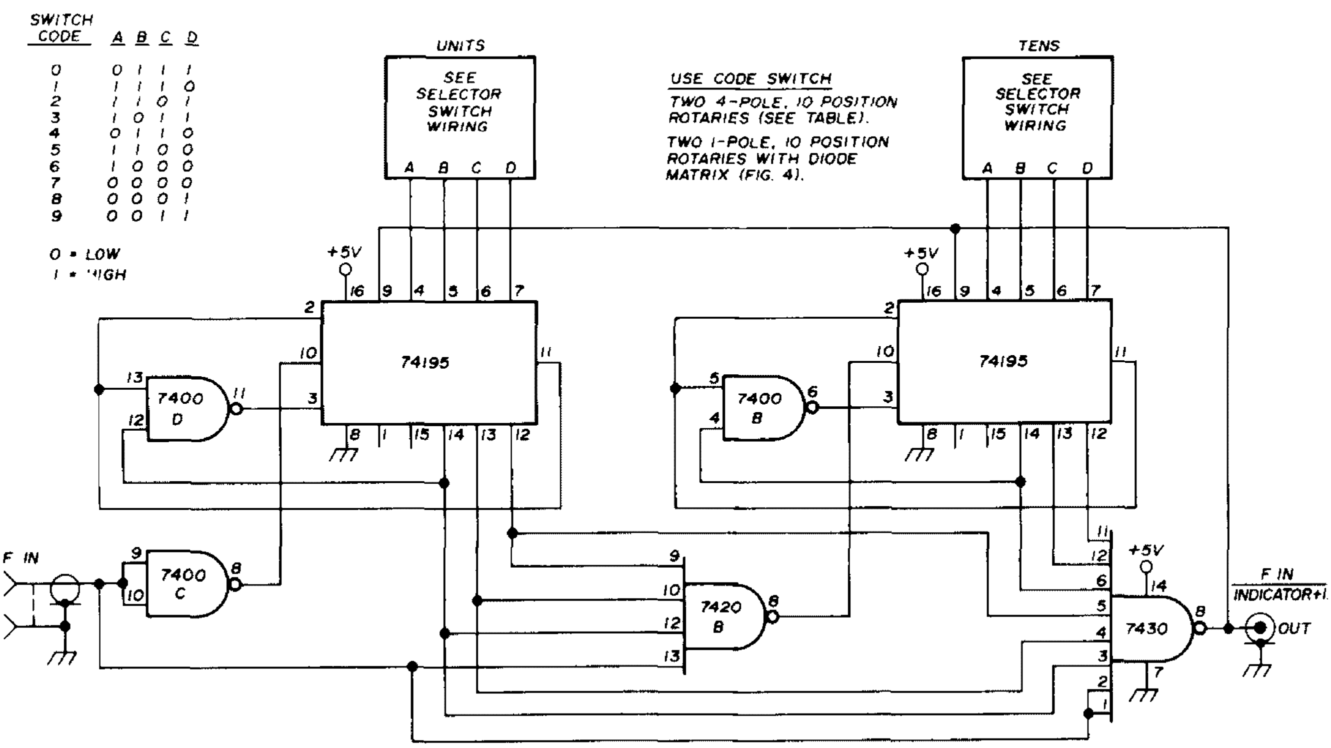


fig. 3. Logic for dividing by any number between 1 and 100. Circuit divides by one more than shown by selector switch (fig. 4), thus dials should be set at 00 to divide by 1; 09 to divide by 10; 29 to divide by 30; etc.

Briefly, the operation of the circuit is as follows. At the end of a divide sequence the registers are loaded with the data from the selector switches connected to wires a, b, c, and d. The registers then clock away until all outputs connected to the 7430 NAND are high, thereby producing an output pulse that loads the data, and the process repeats. The data can be selected by two 4-pole, 10-position selectors wired according to the switch code. A less expensive alternative using diode logic and two 10-position selectors of one pole each is shown in **fig. 4**. The switch code calls for either a 1 or a zero. The

zero means to ground the point, and the 1 means to leave it open. The system counts one more than the data loaded, so if you want to divide by 23 you must set the switches for 22.

The rest of the circuit of **fig. 2** is self explanatory. A part of the 7440 NAND is used to drive a counter. The other part of the 7440 is the mixer, whose output connects to a level-adjusting pot, then to the BNC connector, which leads to the receiver local oscillator. The power for the unit is +5 volts at 225 mA maximum when no crystals are in place. At 15 MHz only 170 mA is required. I used an old 6.3-Vac filament transformer, a diode bridge rectifier, a 1000 μF electrolytic, and a three-terminal, 5-volt regulator. This circuit provides adequate power to the regulator; and at the maximum condition of 225 mA, the supply provides 8.3 volts at the regulator input.

crystal sources

Crystals for the rock-mixer are inexpensive and easy to obtain from a variety of sources. At the Dayton Hamvention, for example, crystals perfect for rock mixing were selling for 15 cents each. The reason the crystals are so inexpensive is because their frequencies are not good for much – except rock-mixing, and you're reading the first account of this now.

Many crystal manufacturers list their surplus crystals at bargain prices, so scan the flyers and you can find lots of "funny-frequency" crystals. Two prime sources I use are the citizens band and surplus fm mobile crystals; for example, some crystals marked for 153 MHz turned out to be 6.38 MHz fundamental mode, and others were 11.6 MHz. In a big bag full of surplus crystals purchased at Dayton were lots of usable ones from 5 to 16 MHz.

connections

Hooking up the rock-mixer to the receiver LO is another of those dealer's choice affairs, because very few amateurs have the same receiver setup. The simplest method, requiring neither holes nor solder, is to run the rock-mixer output coax into the receiver and clip the shield to ground and twist a few inches of the center lead around the wire from the tuning capacitor to the local oscillator coil or tube grid. This

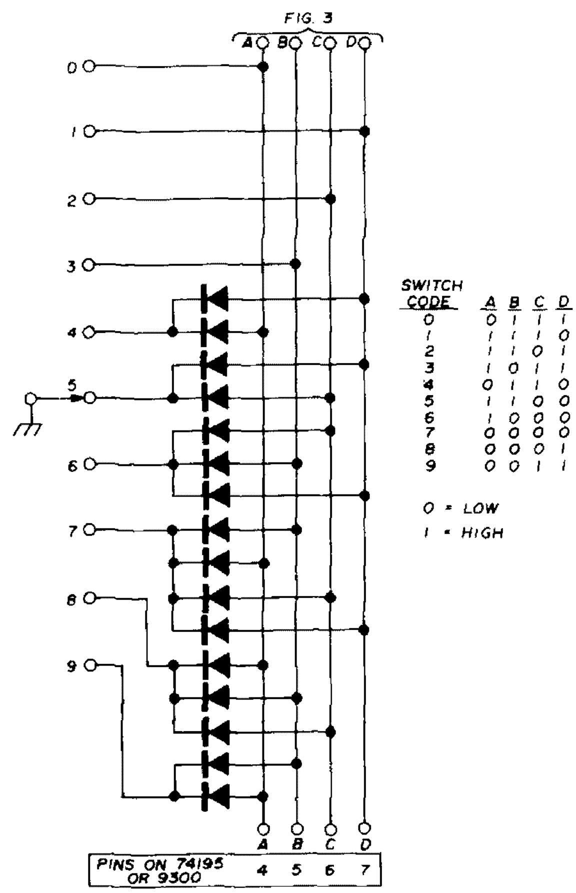


fig. 4. Alternative selector switch and diode matrix circuit.

connection or *gimmick*, injects a weak but significant amount of signal into the oscillator except when the capacitor is nearly meshed.

A better approach is to locate a point in the LO where a ground can be lifted and a 47-ohm resistor inserted. There are many likely spots, such as plate or collector bypass-to-ground capacitors, or emitter or cathode grounds. In the SX-100, for example, the LO has a grounded cathode, tuned-grid, plate-tickler circuit, so in my case a 47-ohm

resistor is now between cathode and ground, and the coax is connected directly to the resistor, terminating in a BNC connector on the front panel. This type of low-impedance injection produces excellent lock-in characteristics over the entire receiver range of 0.5 to 30+ MHz as well as no interference with the normal operation of the receiver. Such a connection is also convenient to bring out a small amount of rf from the oscillator to operate a counter.

concluding remarks

The rock-mixer is a simple device for crystal control of almost any receiver. It's easy to set up and use if you have a counter and a calculator. Its ease of operation will depend on the care used in calibrating both the receiver and VXO. Once learned and recorded, the scheme to lock onto a particular band of frequencies, or even a single point, is quite simple and rapid with or without the calculator-counter combination.

Tuning the band under crystal control is simple if you set up a segment schedule similar to the one shown in **table 2**. However, there's a hitch to this unless you calibrate the VXO to suit the situation, because the receiver dial means almost nothing except to show what band you're on. I solved this problem by building a homebrew counter using decades that can be preset instead of the usual type, which reset only to zero. The counter is switched to read all pertinent frequencies in the rock-mixer and the receiver.

When reading frequency per se, the reset pulse from the logic ties to the normal "reset-to-zero" bus line of all the decades. When connected to the receiver LO, the reset pulse is switched to the strobe data inputs, and the complement of the receiver i-f is thereby loaded into the counter decades. For an i-f of 1650 kHz the complement is 9835.00. Therefore, after 16500 input counts the counters read 0000.00 and

table 4. Receiver lock-in range as a function of division sequence.

received frequency (kHz)	vxo frequency (kHz)	divisor settings shift registers	7493	receiver lock-in range (kHz)
3223	8996	6	2	29.4
3223	8996	6	4	12.2
3223	8996	12	1	5.3
3223	8996	6	1	3.1
3600	5022	11	2	54.1
3600	5022	22	1	4.9
3600	5022	22	2	4.7
8183	9014	11	1	16.0
8183	9014	11	2	12.8
8183	9014	22	2	8.9
8183	9014	22	1	7.9
12175	14977	13	1	38.8
12175	14977	13	2	19.0
12175	14977	13	4	10.0
14030	14998	1	2	59.4
14030	14998	22	1	9.1
14030	14998	22	2	4.1
One case requiring no division:				
8183	9833	1	1	126.0
8183	9833	1	2	88.7
8183	9833	1	4	83.3

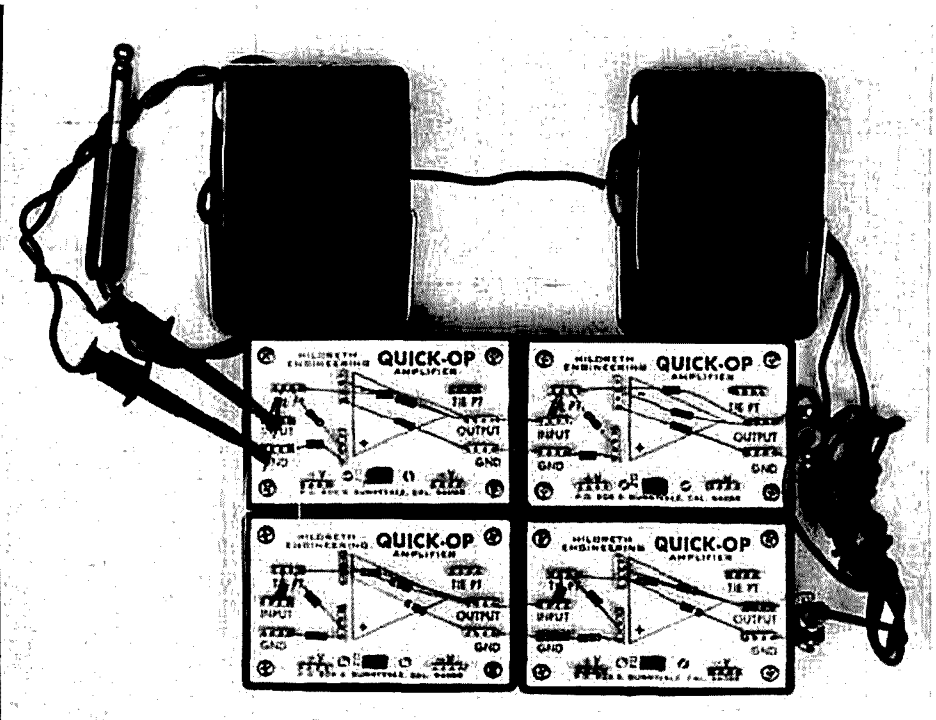
the ensuing input counts above this point until the end of the gate period. The net effect is to subtract the i-f from the LO frequency and to present the received-station frequency in the display.

The action is similar to that of the divide-by-*n* counter, and the data can be either hard wired with short jumpers to ground the appropriate data points; or some can be fixed and some variable, using diode logic or switches made to provide the BCD information. The addition of the data input system in no way affects the normal usefulness of the counter, because the reset input and the data strobe inputs are independent.

references

1. William King, W2LTJ, "Hex Inverter VXO," *ham radio*, April, 1975, page 50.
2. "Multistage Program Divider," *TTL Applications Handbook*, Fairchild Semiconductors, Mountain View, California 94042, August, 1973, pages 9-38.

ham radio



synthesizer for binaural CW reception

Don E. Hildreth, W6NRW, P.O. Box 3, Sunnyvale, California

If you tune your receiver across a CW signal with the bfo centered on a broad i-f passband, the beat note will change from high to low. As you continue tuning, the CW note will progress through zero beat from low to high again.

If, however, instead of using the usual single-channel audio system you provide two channels with the frequency response shown in **fig. 1**, feeding stereo phones or two speakers, an interesting and pleasing result is obtained. Tuning as described, but with the two-channel audio system described here, a signal will move from left (lower frequency) to center, then to the right (higher frequency) followed by the reverse action, spatially as the tone changes. If you switch to a narrower i-f bandwidth and adjust the bfo frequency to equal the crossover point of fig. **1** away from i-f center, you'll obtain the right-center-left signal movement without the mirror image, and the spatial center of the signal will be at the crossover design frequency.

Thus a new dimension is added to CW signals. Can you imagine how it would be when conversing with several people at once if all the voices came from the same location? We have two ears — let's use them.

With the system described, when interference occurs a few hundred Hz or so removed from the frequency of interest, you tune the signal you want to center, leaving the others to the right, or to the left.

components

Op amp active filter designs make this task easy and predictable. As building blocks I used a class of op amp filters shown in fig. 2.* I chose good grade 0.01 μF ceramic capacitors, then selected resistance values from the standard 5 per cent, ¼-watt range to set low- and high-pass filter rolloff fre-

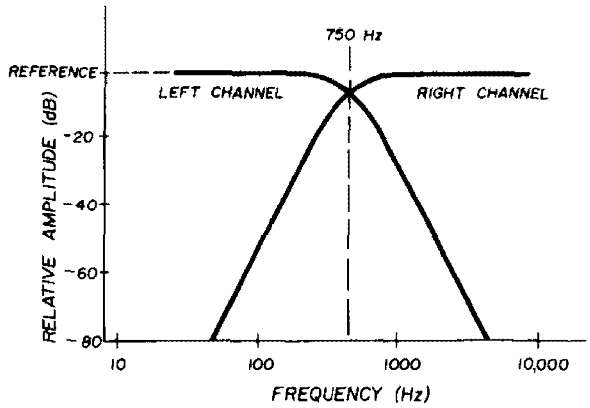


fig. 1. Relative channel frequency responses for right and left binaural CW, voice, or music reception.

quencies. You can use 10 per cent resistors with good results; except that when you try to place the crossover frequency in the center of a narrowband audio filter, or in a very narrowband i-f filter, careful resistance-value pruning enters the picture. I know of no integrated circuit op amp that doesn't loaf in this task – the well-known 741 a good choice.

design requirements

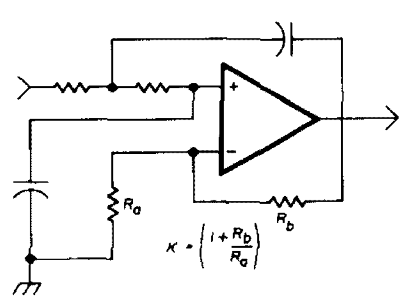
How sharp must the frequency roll-off be to get a good stereo effect? I found that two stages each of low- and high-pass filtering provide good separation (four poles, or a rolloff of 24 dB per octave). More stages and more critical adjustment would be necessary if this system were fed from a source with bandwidth of less than 200 Hz or so. It can be done, but receiver tuning would become difficult and narrowband noise would begin to sound too much like the desired signal. A binaural system reduces the need for those very sharp filters.

The complete circuit is shown in **fig. 3**. You'll note that the gain setting resistors are of different values for the low- and high-pass channels. Their ratio is the same, but values were selected to obtain parallel resistances in each case that are approximately equal to the op amp's positive input port dc resistance to ground. This condition is desirable to encourage a minimum dc offset at the op amp's output.

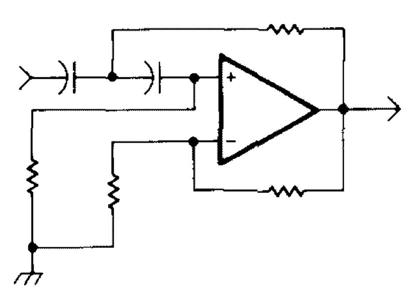
*The op amps used in the binaural synthesizer shown in the facing page are available from Hildreth Engineering, Box 3, Sunnyvale, California 94088. Price is $14.95 each, postpaid. These units use two nine-volt transistor batteries.

exalted operation

If you want to accept some added complexity it's possible to obtain the kind of response shown in **fig. 4**. In this case the gain ratio resistors (those connected to the op amp negative input port) are changed to produce filter peaking. In this way a combination of audio filtering and binaural operation is obtained. The resistance ratio required to provide up to 20 dB of exalted operation is shown (10 dB per stage). It's possible to provide more exaltation effect but adjusting resistance values becomes increasingly critical; at some point you'll have an oscillator.



A LOW-PASS, 2-POLE FILTER



B HIGH-PASS, 2-POLE FILTER

fig. 2. Basic op amp circuits for 3 dB low- and high-pass filters. Center frequency, f_o = $1/2\pi RC$, where network R and C are equal.

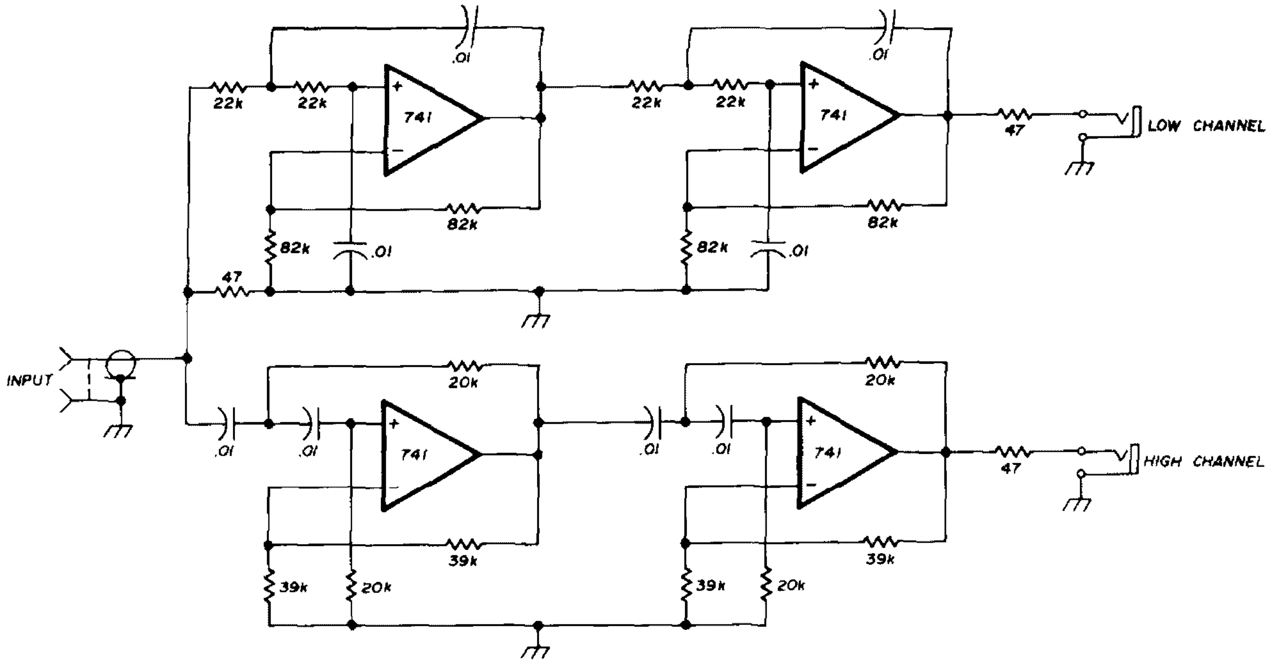


fig. 3. Schematic of the binaural synthesizer. For 750-Hz crossover, R = 21k; low channel is offset 5% high and high channel is offset 5% low to compensate for underlap due to four-pole cascade in each channel and overlap caused by approximately 4 dB of peaking on each side of the crossover point.

The unit shown is designed to work from a low-impedance drive, such as a receiver's speaker output. The 47-ohm resistor at the input is for those who wish to use the design with a solid-state receiver that couples to a speaker through a large capacitor. Op amp input ports must have a dc return path to ground, and this input circuit ensures it. If you want to drive the system from a high-impedance phone circuit, simply connect the receiver to the input through about 2.2k ohms and increase the value of the 47-ohm input resistor to 470 ohms.

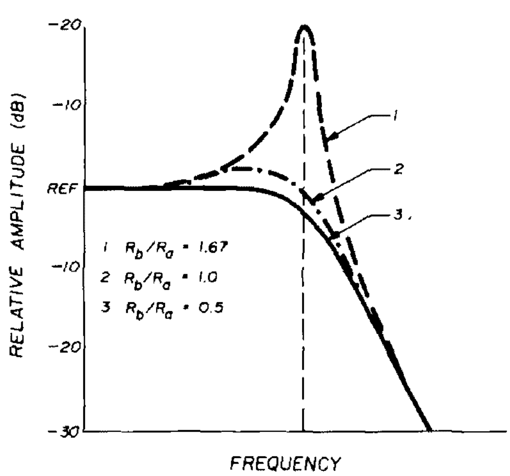


fig. 4. Typical lowpass response shape with peaking adjusted by gain selection of R_b/R_a.

The 47-ohm resistors shown in series with the outputs are used to avoid oscillation in the event that 8-ohm headsets or speakers are connected directly to the output. Also, small dc offsets will not produce immediate limiting on the positive or negative audio half cycle if a low-resistance dc path to ground is connected to the outputs. An op amp's output current capability will drive 2k-ohm phones to all the volume you'll need; unfortunately I haven't found any 2k-ohm stereo phones. Moderate levels are possible with 8-ohm loads.

conclusions

This binaural synthesizer provides more advantage as interference increases. When the band becomes crowded you may wonder how you have been able to get along without such a device.

ham radio

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multiple band

master frequency oscillator

Construction details for a varactor-tuned master frequency oscillator featuring multiple frequency capability

M. A. Chapman, K6SDX, 428 3rd Street, Encinitas, California 92024

The **frequency control** system for any receiver or transceiver being considered for construction is obviously of prime consideration. The voltage-variable diode (varactor) offers significant flexibility in the design and construction of a variable-frequency oscillator. Parallel-plate capacitors suffer from the effects of mechanical rigidity, temperature, humidity, and just plain volume limitations. The varactor is an order of magnitude smaller in size, and is less sensitive to thermal and mechanical stress. For frequency tuning applications, the varactor is considerably simpler to tune and align than its conventional mechanical equivalent.

When a voltage source is used to reverse bias a P-N diode junction, the width of the diode's depletion region varies in proportion to the applied voltage, with the width of the depletion layer increasing with increased bias. This results in an effective capacitance which acts as if it were in parallel with the diode. The amount of capacitance per unit voltage is a function of the variation of the impurity concentration in the depletion region of the diode's P-N junction. If a diode is coupled to a tank coil and the bias across it is varied, the tank circuit resonance will change in proportion to the change in diode capacitance.

circuit design

Fig. 1 shows a precision variable frequency oscillator suitable for general purpose receiver and transmitter frequency control. All rf generating and control components are completely contained within a standard aluminum chassis box.* Since the oscillation frequency is dependent upon the LC ratio and the tuning capacitance is controlled by *a voltage rather than the position* of parallel plates, a potentiometer is used to vary the diode bias between two voltage points corresponding to the desired upper and lower capacitance.

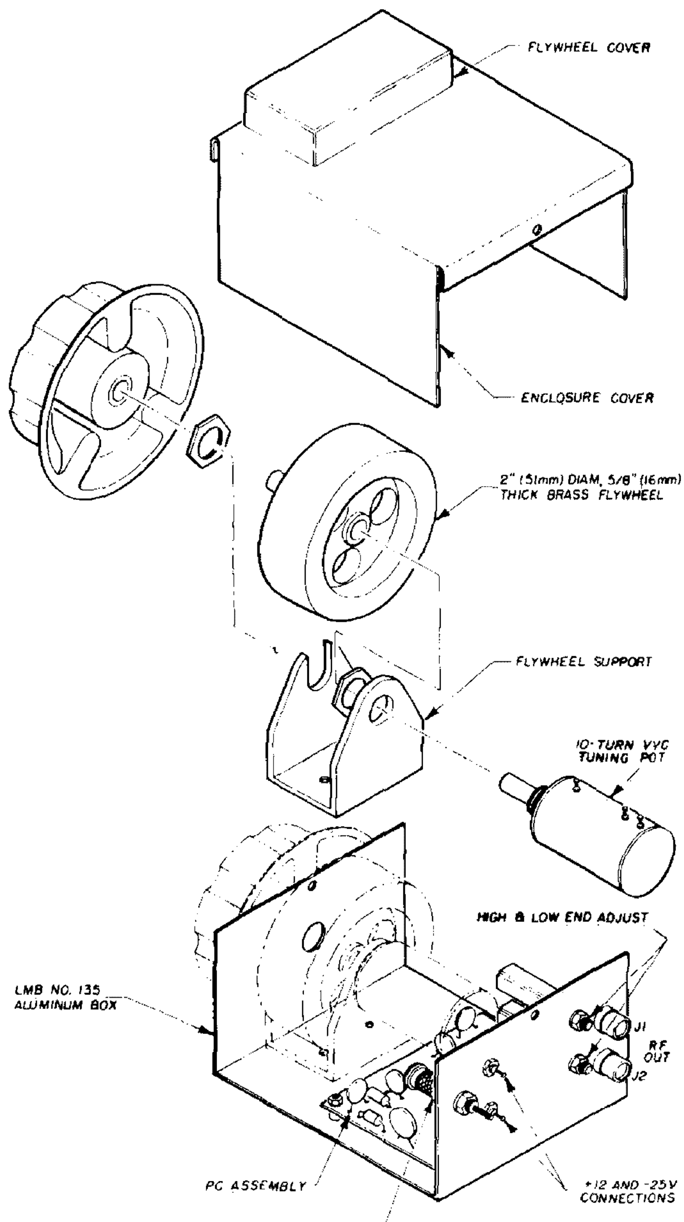


fig. 1. Construction of the varactor tuned master frequency oscillator. Brass flywheel provides very smooth tuning from one end of a band to the other.

If an ordinary single-turn pot is used as the control element, resolution would be very poor, with bandspread similar to a single-turn parallel plate capacitor. To provide additional bandspread in a conventional vfo the frequency-determining capacitor is often mechanically driven by a system of gears so that a single turn of the tuning knob represents a small incremental change in frequency. If a varactor is used as the control element, a small change in voltage will cause a corresponding change in the oscillation frequency.

Although the voltage-control potentiometer for the varactor could be used with a mechanical reduction scheme, most single-turn pots have very poor resolution; at some small discrete points along the wiper surface discontinuities occur which could reflect in an undesirable voltage being applied to the varactor. In addition, many low cost, single-turn pots are often noisy. The noise can be caused by a faulty internal termination, foreign particles or oxidation of the resistance element. The noise then appears as a random voltage, which can cause erratic varactor operation.

If you want to use a single-turn pot in a varactor-tuned vfo, an expensive servo type unit is recommended. The most practical alternative is to use a ten-turn unit, and a wide selection is available at modest cost. With a ten-turn pot to control varactor bias, a full ten rotations of the knob are available for tuning or frequency selection.

All potentiometers have considerable friction associated with the moving element. As shown in **fig. 1**, a simple

*Full-size detail drawings of the chassis are available from the author by sending him a self-addressed, stamped envelope. Etched, single sided, 1/16" printed-circuit boards without holes are also available for $2.00 each, including postage. Write to M. A. Chapman, K6SDX, 428 3rd Street, Encinitas, California 92024.

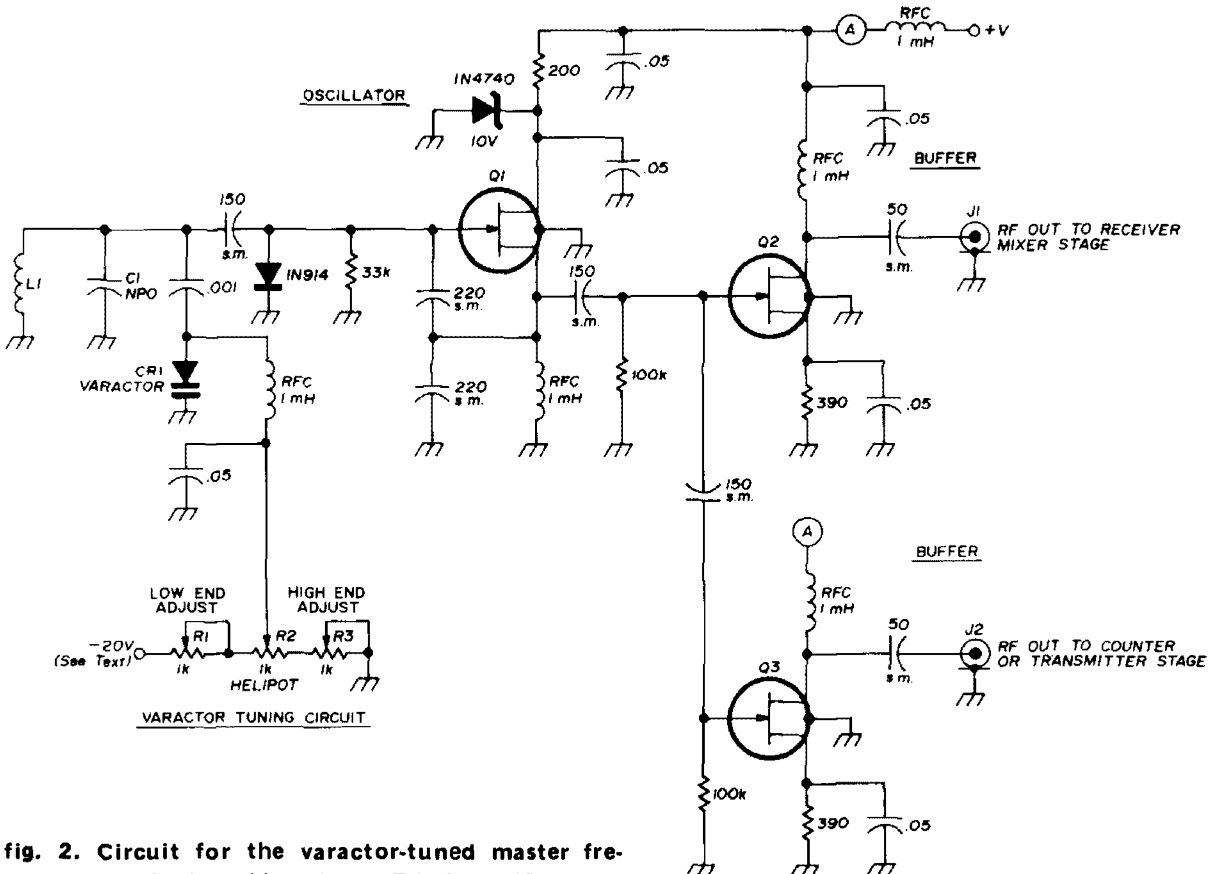


fig. 2. Circuit for the varactor-tuned master frequency oscillator. Varactor CR1 is a Motorola MV1652 or equivalent. Typical tuned circuit values for three popular vfo ranges are listed in table 1. Transistor selection is discussed in the text. R2 is a Beckman model 7426 R1K Helipot; R1 and R3 are Beckman Helitrims, model 78LR1K or equivalent.

flywheel design will provide an exceptionally smooth tuning "feel" for the operator. The suggested 2-inch (50mm) diameter brass flywheel provides enough inertia that a single knob spin will traverse all ten turns of the pot for rapid end to end band tuning.

A circuit for a varactor-tuned master frequency oscillator is shown in **fig. 2** The frequency determining network consists of the inductor, L1, the NPO capacitor C1, and the varactor, A Motorola MV1652.* The 0.001 μF ceramic capacitor provides dc blocking of the varactor control voltage. A 0.005 μF ceramic or similar value will perform as well.

*The Motorola MV1652 is a silicon epicap tuning diode designed for general tuning, trimming and afc applications. Capacitance at −4 volts bias is nominally 120 pF (108 pF minimum, 135 pF maximum). Capacitance ratio from −2 to −20 volts reverse bias is 2.6. The Motorola HEP R2505 closely meets these specifications. **Editor**

The NPO capacitor, C1, can be small as compared to the total change in varactor capacitance for the desired frequency range, and provides some compensation due to thermal effects in the system. Only slight loss in thermal stability would be apparent if a 10 to 30 pF dipped mica capacitor were used.

The frequency range is adjusted by the resistance network consisting of R1, R2 and R3. Maximum varactor capacitance occurs when the *anode* is at ground potential; minimum capacitance occurs when the anode is reverse biased at −20 Vdc. Potentiometers R1 and R3 act as voltage dividers for the main tuning pot, R2.

Current through the varactor is negligible, and scaling of the values shown for R1, R2 and R3 is possible. Transistor Q1 is the basic oscillating element

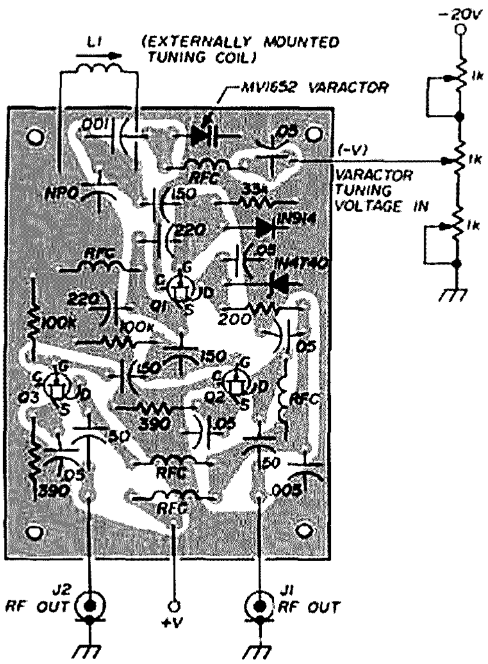


fig. 3. Component placement for the printed-circuit board used for the varactor tuned vfo. Full-sized PC board is shown in fig. 4.

with Q2 and Q3 buffering the output signal to minimize the effects of external loading. The output of Q1 is coupled to both Q2 and Q3 simultaneously so that the buffer stage can drive external circuitry in a receiver or transmitter. The separate buffer stages permit the use of a high level signal from one section in a receiver mixer stage where the conversion gain is dependent upon having an input of 2 volts peak to peak or greater.

The signal from the other buffer stage can be conditioned through filters for transmitter applications requiring lower harmonic content. This is because the normal low-impedance filter will reduce the vfo signal far too much for most receiver mixer applications. The growing popularity of digital frequency displays is also good justification for the separate buffer as it isolates the counter clock from the other receiver circuits.

construction

With the exception of the varactor voltage divider and the inductor, all components are mounted on a printed-circuit board. Fig. 3 shows component placement on the board. All resistors are 1/4-watt units, although 1/2-watt parts may be used by mounting them vertically. Low-value coupling capacitors are of the dipped-mica type; however, glass or silver-mica units are satisfactory. The 0.05 μF ceramic bypass capacitors are low-voltage (20 volt) types; 50 or 100 volt capacitors are approximately the same size and should fit the board equally well. The rf chokes are miniature low-current types.

The selection of Q1, Q2 and Q3 is not difficult. The 2N4416 is the first choice and matches the board layout with case grounding provisions included to minimize random oscillation. The 2N5459 and similar three-terminal epoxy fets work equally well with only slight reduction in output levels. If you have two or three N-type fets in your junk box, give them a try. Even the most general-purpose chopper types I tried seemed to work well.

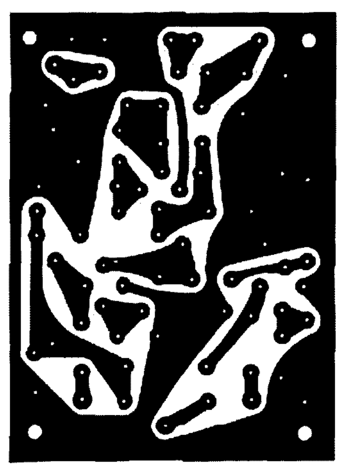


fig. 4. Full-size printed-circuit layout for the master frequency oscillator. Component placement is shown in fig. 3.

Before installing the PC board into the enclosure, attach the wire leads for the inductor, +12 volts and the varactor tuning voltage. The inductor can be temporarily attached to the ends of the wire and left to dangle free in the air. Connect +12 volts and an adjustable negative voltage (not greater than -20

volts) to the appropriate leads. Initial testing of the board can be accomplished in this fashion to insure that the circuit is working properly.

By varying the negative voltage between zero and -20 V, and adjusting the slug in the inductor, a 3 to 6 MHz signal should be present at the outputs. Although the design shown here incorporates the entire circuit in a compact package, the voltage divider and varactor-tuning potentiometer do not have to be adjacent to the varactor. This is one of the advantages of this circuit.

Table 1 lists the LC components and the setting of the frequency-control pot, R2, for three different frequency ranges. Final adjustment of R1 and R3 should not be accomplished until the assembly is installed in the receiver or transceiver because thermal gradients will affect the operating frequency. When adjusting R1 and R3 remember that there is a perceptible voltage change at both ends of the varactor control pot. Inexpensive ten-turn trim pots are recommended for precise adjustment. However, single-turn, low wattage units are satisfactory, although they may require a little more tweeking for final frequency selection.

To calibrate the vfo first set R2 to the appropriate voltage level shown in **table 1** and adjust the slug in L1 for either the high or low end of the band by monitoring the output frequency with a digital counter or calibrated receiver. The output signal level at both J1 and J2 should be between 2 and 3 volts peak-to-peak, depending upon the device used at Q1.

table 1. Tuned-circuit values for three popular vfo tuning ranges. The negative voltage levels shown for potentiometer R2 are for initial setting only; final adjustment must be made in the enclosure to compensate for thermal effects. Varactor is a Motorola MV1652 or equivalent.

frequency range (MHz)	inductor L1	capacitor C1 (NPO)	R2 voltage range (-volts)
3.045-3.545	8.85-12.0 μH Miller 20A105RBI	10-30 pF	3.5-17.0
3.500-4.000	8.85-12.0 μH Miller 20A105RBI	10-30 pF	3.3-12.5
5.000-5.500	≈ 8 μH Miller 20A105RBI (slug removed)	15 pF	5.5-14.0

power supply

A well regulated, low ripple -20 volt varactor bias supply is necessary for best results. The reason for this is apparent if you look at the 3.045 to 3.545 MHz oscillator parameters listed in **table 1**. In this case the ends of potentiometer R2 are at -3.5 and -17 volts, a total range of 13.5 volts. If a 13.5 volt change in varactor bias will produce a 500 kHz change in the operating frequency, a simple calculation will indicate how much frequency variation can be expected for each millivolt of ripple on

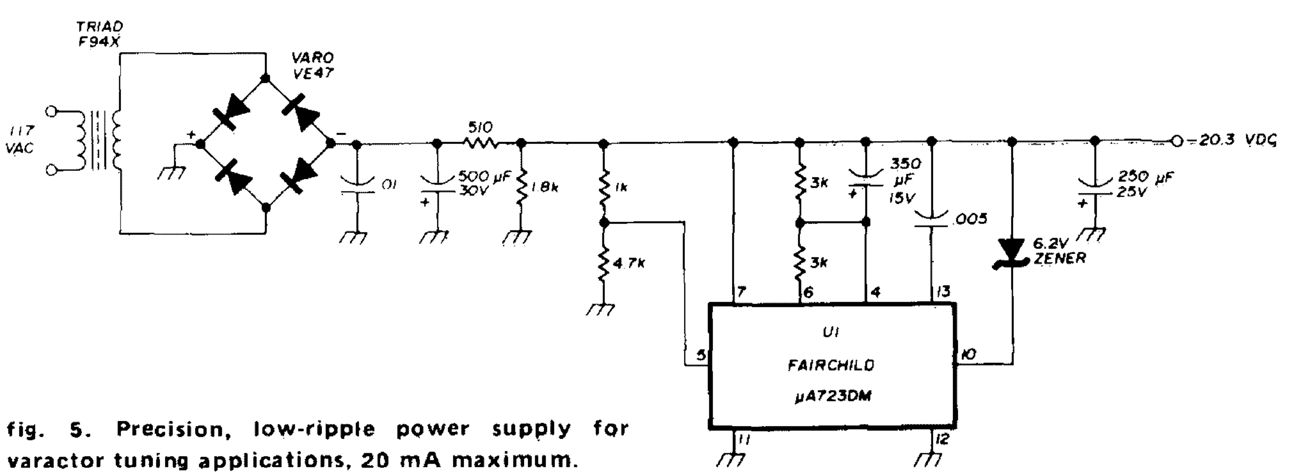


fig. 5. Precision, low-ripple power supply for varactor tuning applications, 20 mA maximum.

the varactor bias supply:

$$\Delta f = \Delta V_B \left(\frac{f_2 - f_1}{V_{B2} - V_{B1}}\right)$$

where Δf is the frequency variation, ΔV_B is the ripple on the bias supply, $(f_2 - f_1)$ is the tuning range, and $(V_{B2} - V_{B1})$ is the change in varactor bias for the tuning range. For the 3.045 to 3.545 MHz vfo

$$\Delta f = 0.001 \left(\frac{500}{13.5}\right) = 37.04 \text{ Hz/mV}$$

Therefore, for each millivolt of ripple on the bias line, there is a corresponding change of about 37 Hz in the oscillation frequency (this would change somewhat at opposite ends of the tuning band). However, the 1 mV ripple across the varactor is related to the ripple on the -20 volt source by the same ratio as the voltage divider. By simple proportion

$$\frac{-20 \text{ volts}}{-13.5 \text{ volts}} = \frac{\Delta V}{1 \text{ mV}} \quad \Delta V \approx 1.5 \text{ mV}$$

For each 1.5 mV of ripple on the -20 volt source you can expect a 37 Hz change in operating frequency. For CW and ssb operation it is desirable to keep the total frequency deviation to less than 150 Hz. This means that the -20 volt bias source should have a ripple content no greater than 6 or 7 mV. This can best be achieved by using precision IC voltage regulators similar to the one shown in **fig. 5**.

parts substitutions

My experience from previous articles indicates that home builders are often faced with parts substitutions, and usually write to the author for advice. A typical case might be the use of a 0.047 μF or other value bypass capacitor as a substitute for 0.05 μF. In this application any value between 0.02 and 0.1 μF should work fine. The 220 and 150 pF units may be replaced with mica capacitors up to approximately 450 pF. The only problem here is board fit, and some capacitors may necessitate some lead bending. The 50 pF mica output capacitors may be replaced with any value from 20 to 500 pF.

ham radio

Yaesu FT101 clarifier

I have just completed the modification to my FT101 as described in *ham radio*[1]. When the modification is completed, depending upon whether the clarifier was turned on or off, the frequency shift may not be a complete zero beat. If the clarifier is turned off when the mod is done, the clarifier will be about 2 kHz high when turned on in the USB mode. If the operator does not care about the calibration of the clarifier, this does not pose a problem, but if he prefers to use the calibration of the clarifier, the following is recommended: Set the clarifier pot to zero before beginning the alignment procedure, and follow the procedure described in *ham radio*. When this is done, and sidebands are changed, the clarifier need not be adjusted to a new zero point and will remain within calibration.

Since I always leave my clarifier turned on, and at the zero position, this is the most comfortable procedure for me. To change sidebands and retune the receiver I just use the clarifier. Otherwise I would have to readjust my thinking when changing sidebands and then turn on the clarifier (as the calibration would not be correct).

Eric Falkof, K1NUN

1. Ernie Schultz, W2MUU, "Yaesu Sideband Switching," *ham radio*, December, 1973, page 56 (short circuit, December, 1974, page 62).

soldering-iron holder

How to build a soldering-iron holder which reduces tip heat when the iron is not in use

E. L. Klein, W2FBW, 137 Ashford Road, Cherry Hill, New Jersey 08003

Perhaps the most important tool used by the electronics experimenter is the soldering iron. It is indispensable when making repairs or building new equipment, and is frequently turned on for hours at a time. This extended time of use takes its toll in corroded tips and burned-out elements.

Radio servicemen learned long ago that they could keep a sharp bright tip on their irons by cutting down the voltage to the iron during those long periods between use. Many old pros would jury-rig a holder on their service bench for this purpose. A bulky heating element was often used in series with the iron when it was at rest in the holder. When the iron was picked up a leaf switch would short out the heating element, allowing full 117 volts to the iron.

In later years commercial holders for soldering irons became available. Some fine thermostatically controlled holders are widely used in the aerospace industry. Printed-circuit boards have called for smaller irons and lower temperatures. Practice has shown that 50 to 70 volts is sufficient to keep the iron ready to go, but low enough to prevent damage to the tip. Variacs have also been widely used to set the iron voltage to the required value.

Described here is a nifty soldering-iron holder that makes use of modern readily available components, and can be assembled in a couple of hours. All parts can be obtained from your hard-

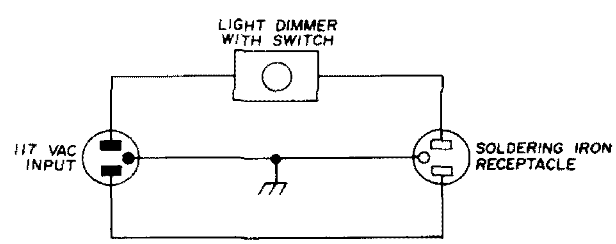


fig. 1. Using a commercial light dimmer to control soldering-iron heat.

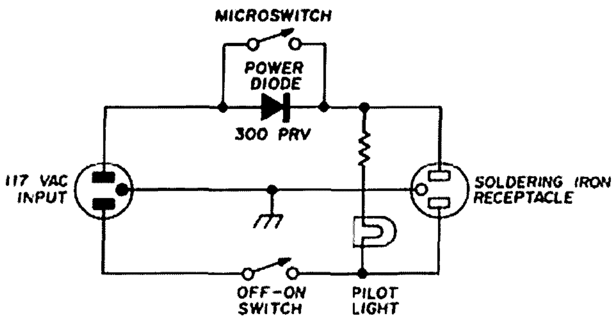


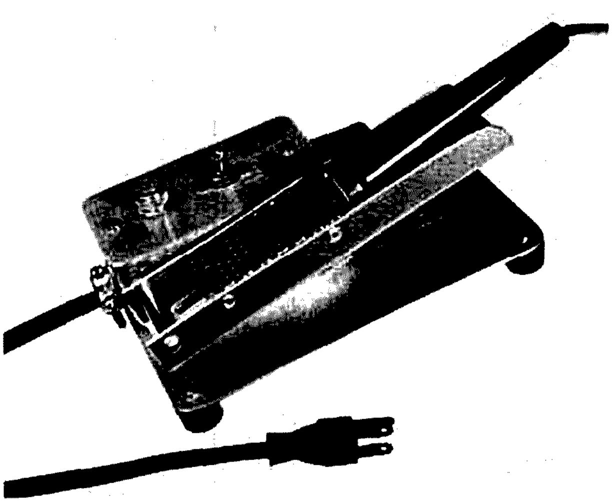
fig. 2. How to use a semiconductor diode to control soldering-iron heat. In this circuit a microswitch shorts out the diode when the soldering iron is lifted for use. When the iron is placed in its holder, the diode is switched into the circuit, reducing the effective power to the iron, by virtue of half-wave rectification.

ware store or radio shop. The sheet metal work is simple and straightforward.

There are two methods for controlling soldering-iron heat. The first simply uses a light dimmer control as shown in fig. 1. Find the position of the knob where the desired heat is obtained at the soldering iron's tip. Then remove the knob. Or, place a mark on the box so the knob can be easily reset when desired.

The other method makes use of a series diode to cut the effective power to the iron. Any power diode with a PRV rating of at least 300 volts will do. The diode is connected across the normally-closed terminals of a microswitch. When the iron is lifted for use, the internal spring in the microswitch operates, returning the switch to its closed position, shorting out the diode. Full power is then available to the iron (see fig. 2). A handy pilot light tells you that power is on and shows the effect of the series diode.

Soldering-iron holder which uses the circuit of fig. 2 to control tip heat when the soldering-iron is not being used.



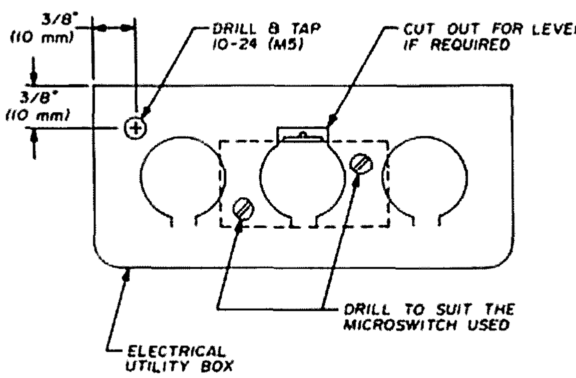


fig. 3. Side view of the utility box which is used to hold the light dimmer (fig. 1) or microswitch/diode circuit (fig. 2). Fig. 4 shows construction details for the soldering-iron holder.

The pilot light shown in fig. 2 is a surplus 28-volt lamp with a 3k, 2-watt series resistor. A neon pilot light is okay if connected to the input, but only one of its two internal elements will glow on rectified ac. If used with the light dimmer, a neon pilot will extinguish at the lower voltage.

construction

First, start with an electrical utility box. Drill and tap for a 10-24 machine screw in the upper lefthand corner as shown in fig. 3. Drill straight through both sides of the box and use a 4-inch (10cm) long screw. This will make a sturdy mount for the holder. A spacer made from 1/4 inch (6.5mm) copper tubing will keep the holder from collapsing when tightening this screw.

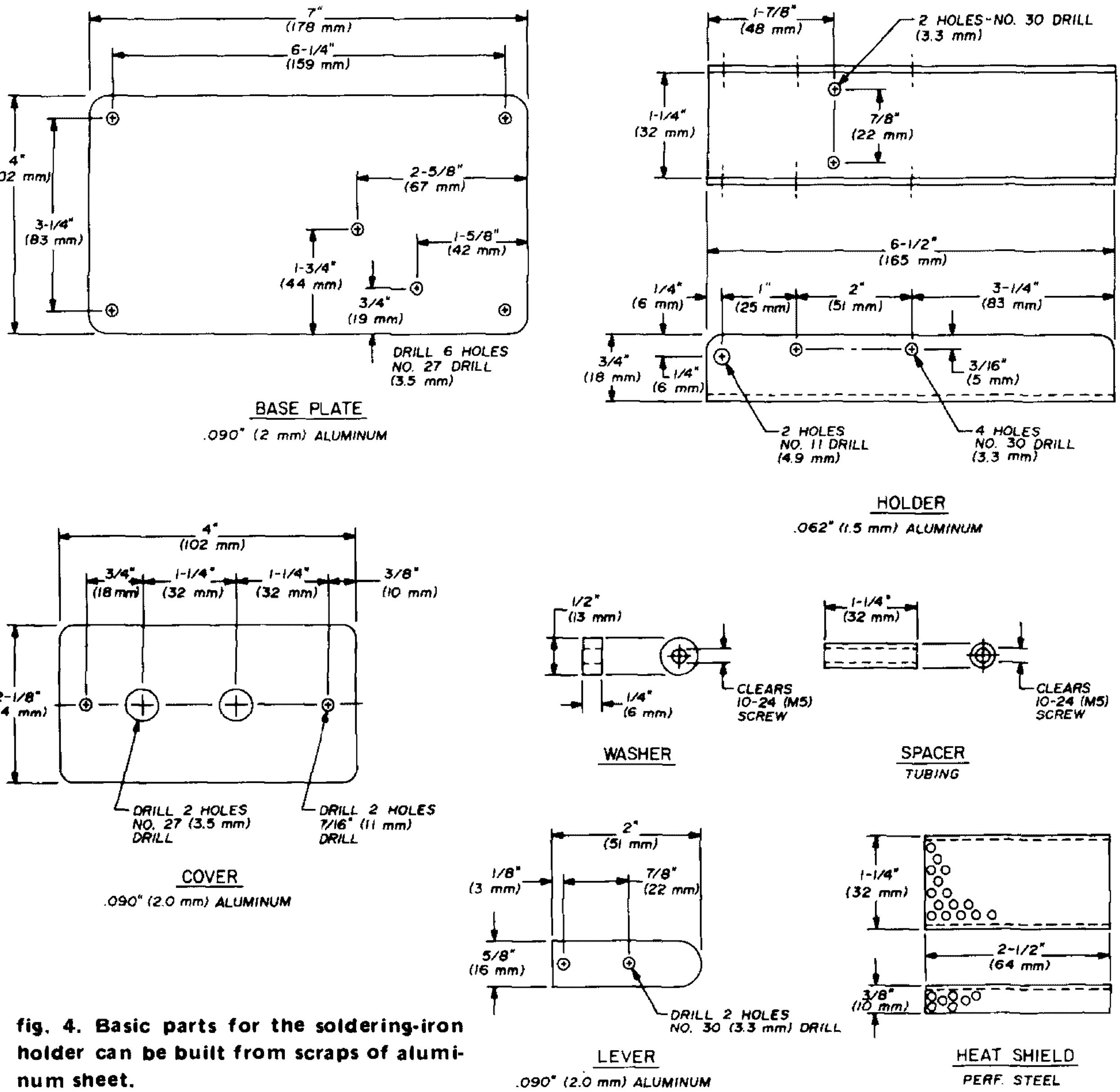


fig. 4. Basic parts for the soldering-iron holder can be built from scraps of aluminum sheet.

Use a conventional wire clamp for the incoming ac line. Choose a socket for the output which has a third wire grounding lug. Use of the ground is important when soldering some ICs and mos semiconductors. Be sure to connect a ground lead from your soldering iron to the circuitry being worked on in these critical applications.

The aluminum parts for the holder can be made from scrap 0.090 and 0.062 inch (2.0 and 1.5mm) aluminum sheet. Other thicknesses can be substituted to satisfy your own design. The heat shield is held in place with four pop-rivets. Two more pop-rivets are used to fasten the lever to the bottom of the iron holder. The lever may have to be shaped slightly to fit through the notch in the side of the utility box so it engages the microswitch. Microswitches are readily available at low cost from many surplus outlets.

Buy a tip cleaner sponge and tray (not shown in the photograph) from a local radio store and cement it on the base just to the left of the ac outlet. Be sure to keep it moistened with a little water. Finally, add rubber feet to the four corners of the base. With your new soldering iron holder you'll have no more burned benches or dull, corroded soldering tips.

ham radio

dipole antennas

A few basic ground rules applied to the installation of this simple antenna can pay off in excellent performance

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The **halfwave dipole antenna** is hard to beat as an effective radiator of rf energy when considered in terms of low cost and ease of construction and tuneup. I'd like to report the results of my experience with this simple antenna for those now using a dipole or who would like to try one. Careful attention to materials, installation, positioning (or orientation), and tuning can make a big difference in performance. Many amateurs swear by the dipole as an antenna for portable work because of its simplicity. The following ideas should be helpful in planning your next dipole or improving your existing installation.

The quarter-wavelength legs of my dipoles have been made of many different materials: insulated copper wire, annunciator wire, aluminum wire, tubing, and even TV twinlead. All will work, but my recommendation is number-14 stranded copper wire. It's easy to handle and causes fewer construction problems than most other materials. Its strength per unit length is excellent and it withstands weather for a long time without failing.

An inexpensive bow and arrow set should be included in your dipole inventory. An 8- or 12-pound (3.6 or 5.4 kg) test nylon fishing line tied to the end of an arrow can be shot over a tree, house roof or similar support. A heavier length of line is then secured to the original line and, in turn, secured to your antenna wire. The whole works is then pulled into position. If the far end of your support is a tree, the arrangement shown in **fig. 1** is one way to eliminate problems with wire breakage due to wind, or wear of the securing line due to friction.

Other antenna supports that can be used are interlocking sections of TV masting, wooden doweling, or the A-frame mast which is described in the *ARRL Handbook*. A husky bamboo pole is another possibility.

installation

I devised the wagonwheel concept (**fig. 2**) to help bring order and logic into resolving the dipole antenna installation problem. Some may think this is a simplistic approach; however, it makes sense to me because any compromise with any segment of the wagonwheel

results in a flat wheel, and who needs that? My dipole wagonwheel has five segments. Let's consider them in order.

Positioning. It might sound trite, but the best antenna is one that's located as high as possible and in the clear. This means the radiating (and receiving) wire should be positioned as far as possible from telephone wires, metal house siding, fences, and the like. If the antenna is located close to trees or shrubbery,

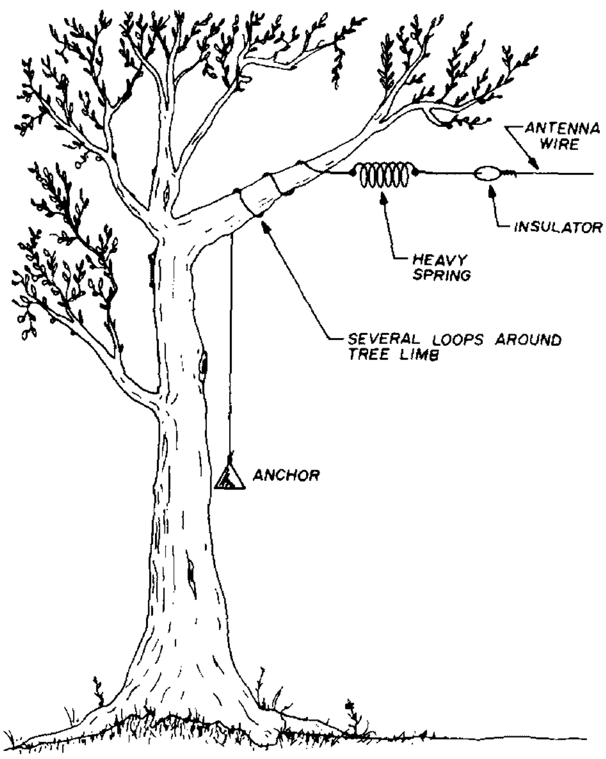

fig. 1. Suggested method of securing the far end (or both ends) of a dipole antenna, using nature's remedy.

the electrical characteristic of the reflecting surface will be adversely affected.[1]

Each of my dipoles is constructed for one band; ideally the height above ground for each antenna should be one-quarter wavelength minimum for that band. Notice I said "ideally." The ideal situation is hard to achieve. If you must compromise on antenna height, try to compensate by observing the other installation hints mentioned here. See **fig. 2**.

Another consideration is the placement of two or more antennas. I once tested a long-wire antenna that had an antenna tuner and an swr meter in the transmission line. i was transmitting using a dipole about 15-feet (4.58m) away. It turned out that the long-wire antenna was absorbing a great portion of the signal radiated from the dipole. This makes me wonder how much power is lost in direct and harmonic absorption. So now my rule is, "Keep antennas separated and preferably oriented in a different plane of transmission."

Resonant frequency. This is the second segment in the dipole wagonwheel (**fig. 2**). After selecting your desired band and the part of that band in which you wish to work, the leg lengths of your dipole are easily determined from formulas in the *ARRL Handbook*. The lengths given in these formulas are usually somewhat long, which is fine for cut-and-try installation.

The resonant frequency is most accurately determined when measurements are made as close as possible to the base of your antenna. You'll need an rf source and, depending on the technique you wish to use, an swr meter, grid-dip meter/antenna bridge, or noise bridge.

The swr technique is easiest to use in testing a dipole antenna. However, there are restrictions as to the readings because there is no direct method of exactly reading either frequency or impedance. After the dipole has been placed in its operating position, a test length of coax cable is attached to the feedpoint, which should be a balun (see the discussion on wagonwheel segment 4). The other end of the test coax line is attached to your rf source (a transceiver in my case).

My test piece of coax is cut for a multiple of one-half electrical wavelength at the frequency at which I wish to test my antenna. I cut the coax test line slightly longer than a multiple of

one-half wavelength, then made a shorting device from a straightened safety pin to obtain this length exactly. This test line can be used for all three methods of antenna testing.

With the test line attached to the swr meter and the rf generator, tune across the band in 100-kHz increments with the set tuned to maximum output, then

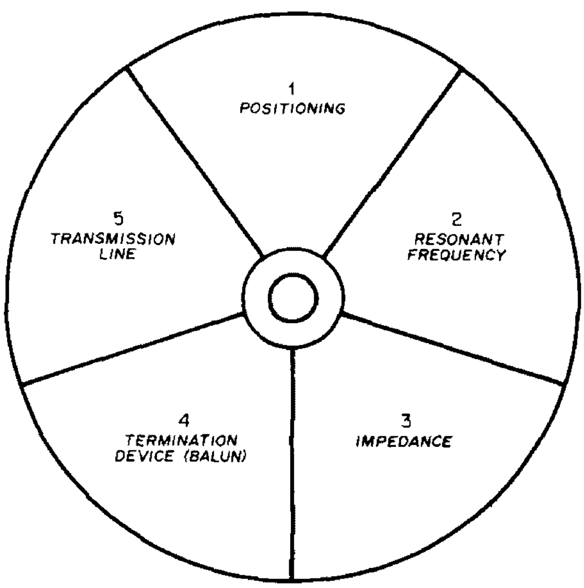

fig. 2. The dipole wagonwheel — a useful adjunct to the installation problem.

reduce power. The swr meter is set at full forward reading, then reflected power is recorded. Where the lowest reflected-power reading occurs is the antenna resonant frequency. If the swr is more than unity, don't worry. An swr of 3 or 4 is acceptable in amateur work.

Impedance. Using the grid-dip meter/antenna bridge method, select the correct frequency probe for the grid-dip meter, and with the meter turned on, you can spot your desired frequency on the transceiver. The grid-dip meter acts as a transmitter for your desired frequency. Connect the test line to the antenna bridge. By varying the grid-dip meter dial, you'll get a dip on the antenna bridge at the antenna resonant frequency and you'll know whether the wire is too long or too short. The antenna bridge will also show antenna impedance.

In the noise bridge method, an a-m receiver signal is used. The bridge is attached to the receiver and turned on. The noise bridge will produce an output that is like atmospheric noise. As you tune the receiver across the band for which your antenna is cut, you'll obtain a null in the noise at the antenna resonant frequency. As with the antenna bridge, the noise bridge has an impedance dial that, when set to your antenna impedance, will produce a noise null. You can read an antenna resonant frequency and *impedance fairly* accurately.

At this point I'd like to include two *personal notes. First, the* bridge operates only with an older type a-m receiver. Second, be sure your leads from the receiver to the bridge are short (not over 10 to 12 inches or 25.4 - 30.5 cm). I made these notes not from textbook directions but from yardwork failures.

Antenna impedance matching can fill a large textbook. With a dipole you can approach 50 ohms by changing the angle of the legs from the horizontal or by using a matching system. With my 20-meter dipoles, I use a piece of wire 42 inches (1.07) long with a clip at each end. One clip goes to each side of the

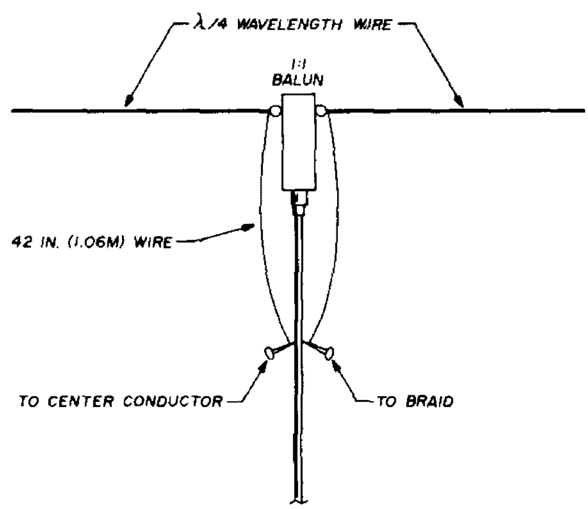

fig. 3. Delta match. Pins are used to obtain correct impedance match. Permanent installation should be soldered and weatherproofed.

balun connection (fig. **3**). This matching system works well for me. An alternative is a matching stub as shown in fig. **4**.

Balun. For quite a time I didn't understand about the balun and therefore didn't use it. Later I used the balun incorrectly thinking that it tuned out all of my antenna faults. However, the balun is

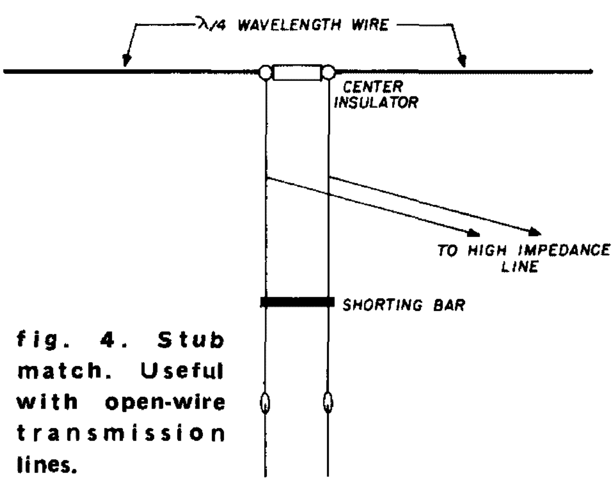

fig. 4. Stub match. Useful with open-wire transmission lines.

a necessary device in a balanced antenna system such as a dipole. Radio-frequency energy propagates along the coax at a different rate in the shield compared to that in the center conductor. The result is that antenna currents will appear on the outside of the shield braid, and these currents will radiate.[2] Such radiation causes undesirable antenna performance and is often the culprit in TVI. The balun will often reduce rf radiation from the coax.

Transmission line. Use only top-quality foam-dielectric RG-8A/U coax cable. I have lengths of cable to reach from the antenna to my swr meter, which are cut in multiples of one-half electrical wavelength determined from charts in coax cable handbooks and checked with a grounding pin made by straightening a safetypin. Using the grid-dip meter and antenna bridge test described earlier will produce such a length of coax.

In testing the correct length of coax to use, start at the signal source with the antenna bridge/grid-dip meter setup and include swr meter, in-line wattmeter, and lowpass filter. Coax transmission line lengths can also be measured in one-half electrical wavelengths by using an antenna noise bridge.

dipole variations

The dipole is the basis of all high-frequency antenna systems. This simple structure can be expanded almost without limit to produce extremely complex directive arrays. For example, a directive antenna used in France in the early 1940s for shortwave transmitters operating around 8 MHz had three horizontally oriented bays of six dipoles each, with each set of dipoles arranged in a diamond configuration. Each bay of six dipoles constituted one element of a three-element beam antenna — director, driven element, and reflector. This system was known as a Chireix-Mesny array.[3]

The currents along any diagonal of each diamond had to be exactly in phase, so that the antenna wires served both as radiators and transmission line — truly an installer's nightmare. Such arrangements were popular for a short time but were eventually abandoned because of construction expense and tuning difficulties.

I have installed wire directors and reflectors in my dipole systems, used coil traps, and tried to decrease physical space requirements by installing the dipole ends at different angles from the horizontal. I still like the simple half-wavelength dipole as described and built according to the wagonwheel concept shown in **fig. 2**. This is the antenna I use today — simple and effective.

references

1. Arnold B. Bailey, *TV and Other Receiving Antennas*, pp. 182 - 183, John F. Rider, Inc., New York.
2. *The Radio Amateur's Handbook*, ARRL, Newington, Connecticut.
3. F.E. Terman, *Radio Engineers' Handbook*, McGraw-Hill, New York, 1943.

ham radio

Collins R390A modifications

Several simple modifications for the R-390A which can considerably improve performance

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With the R390A receiver, it sometimes pays to work around a built-in problem or known trouble, rather than make extensive repairs. Here are some simple and somewhat complex modifications that may help, depending on the trouble you have.

Audio section. If you have audio problems, it may be due to a mismatch. All of the audio outputs are 600 ohm. A pair of high-impedance (3000 ohm) headphones will work as is. For low-impedance (8 ohm) speaker or phone use, the output transformer from a small tube type receiver or one of the universal replacement types will provide a reasonable match.

If there is still trouble, the entire audio section can be completely bypassed by using the diode load terminal at the back of the set. Leave the jumper connected and couple the signal to an outboard amplifier and speaker through a suitable blocking capacitor (fig. 1). A hi-fi amplifier used with the R-390A will give you beautiful shortwave broadcast-band listening. The added clarity will help amateur band reception too.

I-f section. Ssb reception with the R-390A has a mushy audio quality because of the envelope detector and the low bfo-to-signal ratio. One solution is to rewire the detector as a product detector.[1]

The set can also be used with a companion ssb adapter fed by the i-f output jack. There are advantages to building an adapter for the set rather than converting the existing circuit: you will have more room to work with; you can choose your own parts layout; and, you can build as elaborately as you want. More important, you should be able to get better performance from a totally outboard unit than by piece-meal modifications to the set.

Originally these sets were stagger tuned to improve the bandwidth characteristic. For amateur use the i-f stages can be retuned to the same frequency, which noticeably increases gain.

Rf section. When you increase the gain you also increase the noise. The rf gain control works in both the rf and i-f sections. As the set is now there is no way to vary the i-f gain without adversely affecting the rf stage.

The rf stage determines the overall sensitivity and noise level of the set. Removing the rf amplifier cathode circuit from the rf gain control and grounding it directly lets the stage work at its maximum gain and sensitivity. The cathode resistor (**fig. 2**) runs from the tube socket to a nearby standoff insulator where it connects with the rf gain control wiring.

Remove the wire from the standoff and tape it out of the way so it can't short. Then run a short wire from the resistor end on the standoff to a convenient ground lug. It would be a bit fussy, but you could run a shielded cable to a switch on the front panel and make the modification optional.

While the modifications are simple to make, you will need the manual to safely disassemble and reassemble the rf deck. Without it, it is too easy to damage the set or misalign the tuning mechanism.

It is possible to position the rf deck on its side in the main chassis so that the cables will just reach and you will be able to get at the bottom of the subsection chassis for testing or trying modifications with the set in operation.

Be careful when doing this as there is almost no slack in the cables and it is

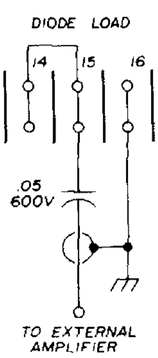

fig. 1. Diode load audio tap for use with an external audio amplifier.

very easy to break one or damage some other part. Replacing one of the coax cables would try the patience of a saint.

Low sensitivity. If your R-390 seems to have lower sensitivity and higher noise below 8 MHz, and no fault can be found, try bridging either C281 or C282 (first mixer output coupling capacitors) with a higher value; the gain may come right up. The value probably isn't critical; I replaced both capacitors with 100 pF.

Antenna matching. The unbalanced antenna jack, J103, was intended for a

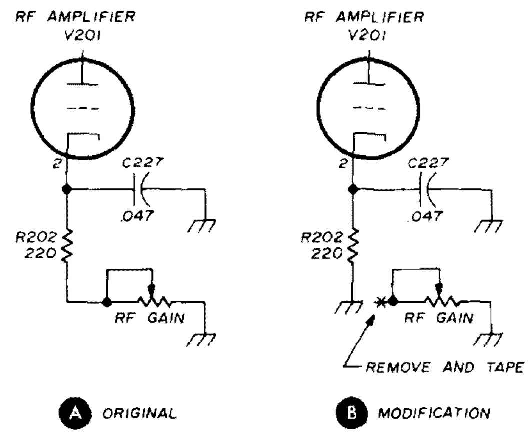

fig. 2. Modifications for the R390A rf amplifier.

whip antenna with a very short lead-in or a random length of wire. If you are using a longer length of coaxial cable you may be losing most of the signal.

A UG-970/U adapter, used with balanced input jack J104, makes the necessary changes with a substantial improvement. The following modification, originally issued by the Navy as a field change, does much the same thing.

1. Disconnect plugs P205 and P206 from the antenna box inside the set and reverse them: P205 to J106 and P206 to J105.

2. Connect a shorting plug to J104.

3. Connect the antenna to J103 which, because of the internal changes, provides a much better match to the antenna.

reference

1. Eugene A. Hubbell, W7DI, "Improving the R390A Product Detector," *ham radio*, July, 1974, page 12.

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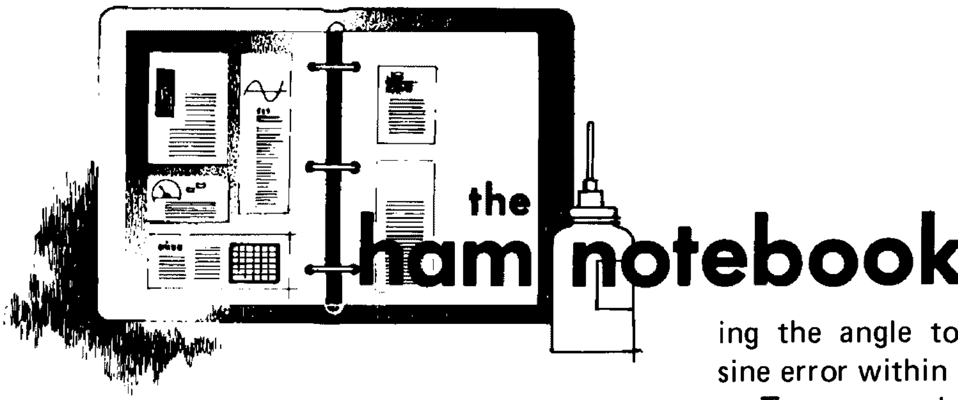

trig functions on a pocket calculator

There are several ways of evaluating log, exponential and trig functions on small hand-held calculators. Here is a method for trig functions which offers some advantages if the calculator has square-root capability. Methods for finding square roots on four-function machines were previously described in *ham radio*.[1]

The usual scheme is to run out the calculations using the series expansion for the sine or cosine. For the simple four-function machine this has the advantage of requiring only four basic operations. However, it has some disadvantages. The infinite series expressions are difficult to remember. Also, a number of terms of the series must be added together to arrive at a value accurate to three or four decimal places.

With square-root capability sine, cosine and tangent can be done quite simply by making use of a few trig identities. A useful approximation is that for small angles, the sine, the tangent and the angle, expressed in radians, are equal. Table 1 lists some values along with the error for using the angle (in radians) rather than the sine or tangent function. Note that the values for the sine are somewhat closer to the actual values than for the tangent — about a two to one difference. Up to 20 degrees the maximum error is 4 per cent; limiting the angle to 15 degrees keeps the sine error within 1 per cent.

To convert degrees to radians simply multiply the angle by pi and divide by 180. If the angle is 15 degrees or less this immediately gives the approximate sine or tangent. To evaluate the cosine use relation (3) after obtaining the sine. This is very simple on a calculator with square-root capability.

For angles between 15 and 45 degrees use relation (5). First calculate the sine and cosine for an angle that is half the desired angle. Then multiply these two together and times 2 to obtain the sine of the angle. Memory is useful during this double calculation to store the intermediate value for the sine.

To increase the accuracy of the re-

table 1. Values for the tangent and sine of small angles are very close to the angle expressed in radians, as shown here. Trig identities for calculating other functions are shown below.

θ (degrees)	θ (radians)	$\sin\theta$	error	$\tan\theta$	error
1°	.01745	.0175	0	.0175	0
2°	.03491	.0349	0	.0349	0
5°	.08727	.0872	0.1%	.0875	0.3%
10°	.17453	.1736	0.5%	.1763	1.0%
15°	.26180	.2588	1.2%	.2679	2.3%
20°	.34906	.3420	2.1%	.3640	4.1%

$$\theta(\mathrm{RAD}) = \frac{\pi}{180} \times \theta(\mathrm{DEG}) \qquad (1)$$

$$\mathrm{FOR}\ 0 \leqslant \theta \leqslant 15^\circ \quad \sin\theta \approx \tan\theta \approx \theta(\mathrm{RAD}) \qquad (2)$$

$$\cos\theta = \sqrt{1 - \sin^2\theta} \qquad (3)$$

$$\tan\theta = \frac{\sin\theta}{\cos\theta} \qquad (4)$$

$$\sin 2\theta = 2\sin\theta\cos\theta \qquad (5)$$

$$\sin\theta = \cos(90^\circ - \theta) \qquad (6)$$

1. John Sego, K9DHD, "Finding Square Roots," *ham radio*, September, 1973, page 67.

sult, the above procedure is done in two steps at angles one quarter and one half the desired angle. Between 30 and 45 degrees this method is almost mandatory since the error above 15 degrees is fairly large. An example, **table 2**, has been worked out for the sine of 45 degrees. The calculated value differs by only 0.5 percent from the actual value.

For angles between 45 and 90 degrees use relation (6). Find the sine and then the cosine of the complement of the angle desired.

table 2. Using the trig identities shown in table 1 to calculate the sine of 45 degrees. Steps can be accomplished easily on a pocket calculator with square-root capability.

$$\sin 22.5^\circ = 2(\sin 11.25^\circ)(\cos 11.25^\circ)$$

$$11.25^\circ = \frac{\pi \times 11.25}{180} \text{ rad} = 0.19635 \approx \sin 11.25^\circ$$

$$\cos 11.25^\circ = \sqrt{1 - (0.19635)^2} = 0.98053$$

$$\sin 22.5^\circ = 2 \times 0.19635 \times 0.98053 = 0.38505$$

$$\sin 45^\circ = 2(\sin 22.5^\circ)(\cos 22.5^\circ)$$

$$\cos 22.5^\circ = \sqrt{1 - (.38505)^2} = 0.92289$$

$$\sin 45^\circ = 2(0.38505)(0.92289) = 0.71073$$

$$\sin 45^\circ \text{ (from trig table)} = 0.70711$$

$$\text{error} = \frac{0.71073 - 0.70711}{0.70711} \times 100\% = 0.512\%$$

As indicated in **table 1**, the error for the tangent is somewhat larger than for the sine when using the angle in place of the function. Since the above steps provide simple calculations for both the sine and cosine, relation (4) can be used to find the tangent of any angle.

The trig identities shown here should be at least as familiar as the series expansions for sine, cosine and tangent. In fact, relation (5) is really the only special identity in the group; the others come from trigonometry definitions.

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copper-plated circuit boards with terminal inserts

Perfboard with terminal inserts has served well for many projects. What it lacks is the all-important ground plane that an etched board provides. This ground plane can be the difference between a quiet and a noisy mike preamp or the difference between a smoothly acting rf or converter stage and one that has a will of its own.

The answer I developed is a marriage of a circuit board copper plated on one side only with the perf board insert terminals. Insulated islands for the terminals was the immediate problem. The solution for this was to use a bit designed to rout channels in wood. Chucked in the drill press, this routing bit takes perfect ¼=inch (6.5mm) circles of copper from the board. Holes are then drilled in the center of the newly created insulated islands and the perf board terminals are inserted. The circuit is then wired point-to-point, with any components requiring a ground being terminated in a hole in the copper ground plane and soldered directly to the copper.

Layout is a common sense approach. Merely pencil a grid on the copper surface, determine where you want the islands, and apply the router bit. A few moments practice on a scrap piece of board will quickly give you the feel of just how much pressure to apply with the router bit to get perfect removal of copper without biting into the board proper.

Duplicate boards may be made by applying identical grids to the blanks. After the islands have been created, the boards may be stacked and drilled in one operation for terminal insertion. If you use care, this method may be used with board material plated on both sides.

Allan S. Joffe, W3KBM

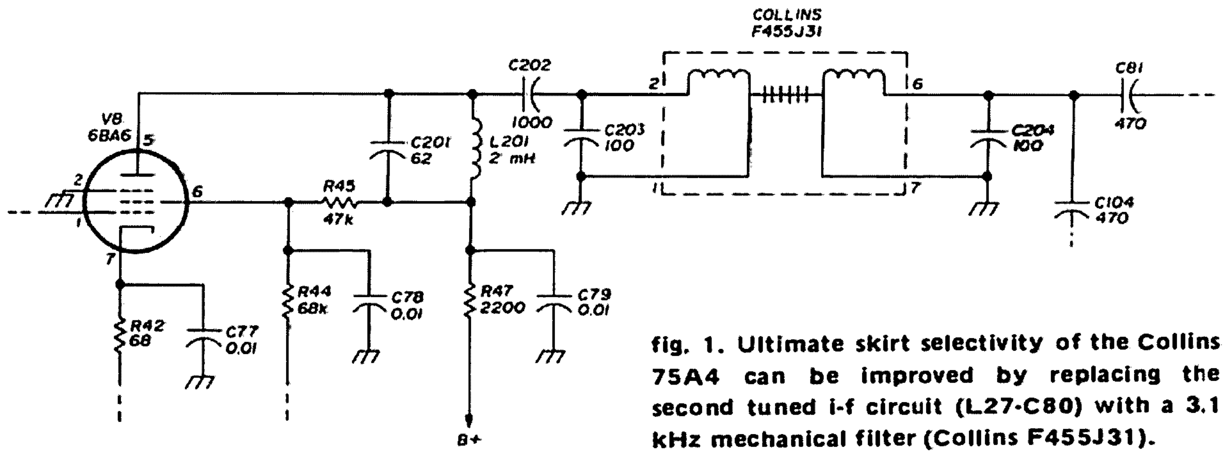


fig. 1. Ultimate skirt selectivity of the Collins 75A4 can be improved by replacing the second tuned i-f circuit (L27-C80) with a 3.1 kHz mechanical filter (Collins F455J31).

headphone cords

For some time I have tried to purchase replacement earphone cords for my headphones. Over one dozen New York merchants told me they didn't stock them.

Various alternatives (including four-wire rotator cable) were tried, but none of them were satisfactory. If you are faced with a similar problem I would suggest trying Trimm, Inc., for suitable replacement cords.

I tried both their no. 811, standard pin tip terminals, black cotton braid, 4½ feet (1.4m) long; and their no. 870, similar but 5 feet (1.5m) long with a waterproof outer braid. Costs range from $2.00 plus postage. A card to Trimm, Inc., Post Office Box 489, Libertyviile, Illinois 60048 could save you a lot of exasperation.

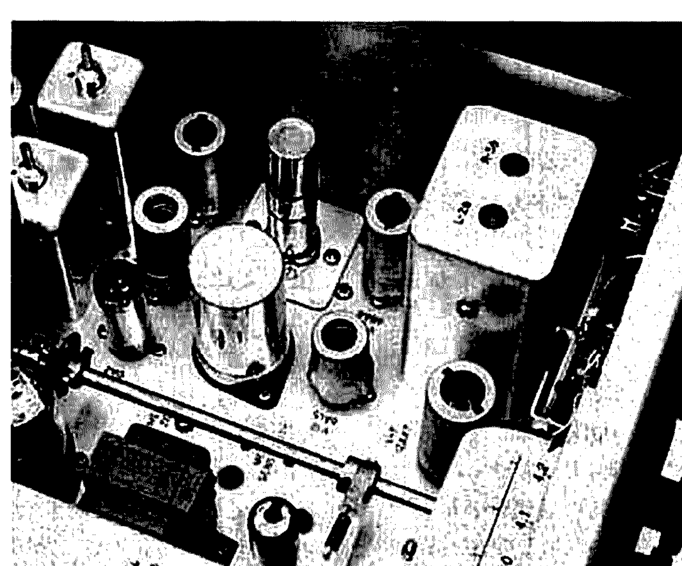
Neil Johnson, W2OLU

increased selectivity for the Collins 75A4

The ultimate skirt response of the 75A4 selectivity curve can be improved considerably by replacing the second 455-kHz i-f amplifier tank circuit (L27-C80) with a 3.1-kHz Collins mechanical filter (F455J31) as shown in fig. 1 (modification suggested by W4ZKI). Since most amateurs who use the 75A4 for ssb operation have replaced the original 3.1-kHz filter with a 2.1-kHz filter, the 3.1-kHz unit is seldom used. If a 3.1-kHz filter is not available, a 4.0 kHz filter (F455J40) will still provide a noticeable improvement in skirt response. The L27-C80 tuned circuit is in the i-f can next to the filter capacitor, C94.

Remove the bottom panel of the receiver, disconnect all the leads which go to the L27 i-f can, and remove the two retaining nuts (don't discard the i-f can — you may want to restore the receiver in the future). Cut out a small piece of thin aluminum, 1-3/4 inch (4.4cm) square, and punch a 3/4-inch (2cm) hole in the center for a 9-pin tube socket. Drill the two chassis-mounting holes and position the tube socket so pins 1-2 and 6-7 are aligned with them. Install the

Filter installation in the Collins 75A4.



socket on the plate and fabricate a small brass shield about 5/8 inch (1.6cm) high. This shield is placed across the tube socket between pins 3-4 and 8-9 and soldered in place (see mechanical filter sockets A, B and C for reference). Ground all unused socket pins.

Wiring the new filter into the circuit is straightforward and requires only four mica capacitors and one inductor. (C201-C204 and L201 in fig. 1). Install the two 100 pF filter resonator capacitors at the input and output socket pins (the filter is symmertical so either set of pins may be used as the input). Install a small terminal strip next to V8 for the junction of R45, R47, C70, C210 and L201. Delete C69 and R46 as they are not used in the new mechanical filter circuit.

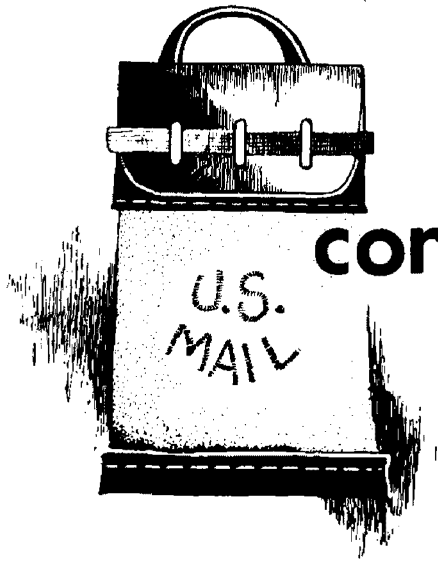
An improvement in i-f gain can be obtained by removing resistor R29 from the plate circuit of V6. This resistor swamps out the Q of L24 and increases the bandwidth for a-m reception; it is not required for ssb or CW operation.

Jim Fisk, W1DTY

muting microphones

Other amateurs must be faced occasionally with the same problem I was: that of disturbing others in the household when talking into a microphone. Headphones, of course, eliminate any speaker disturbance. The microphone problem was solved by attaching a heavy-walled cardboard tube (of the proper diameter) about 3 inches (10cm) long to the face of the microphone, making sure the joint is completely sealed. By pressing your lips into the open end of the tube, and speaking in a whisper, no sound can be heard in the shack. The fact that the voice is completely retained within the tube compresses the sound, resulting in increased talk power, although it may sound like you're in a barrel. Microphone gain must be reduced considerably.

Ralph Cabanillas, Jr., W6IL



comments

speech processing

Dear HR:

ZL1BN's article on speech processing in the February, 1975, issue of *ham radio* covered much of the knowledge which has been available in the literature to the amateur. However, a few ideas to which I have been exposed were omitted. First, no reference was made to the excellent article in *QST*[1] which developed a theoretical and empirical model of *intelligibility*, a concept which continues to be confusing to most hams with whom I have had relevant discussions.

Secondly, ZL1BN is well justified in his concern for problems of signal-to-noise degradation and power inefficiency resulting from heavy clipping levels, as can be attested by those who have heard, say, a Signal One under full clipping and power in heavy competition. One method which I have found effective in reducing extraneous noise in conjunction with my rf clipping system is the use of an "inverted" audio compressor; that is, an expander. A moderate amount of expansion (5 to 10 dB) seems to keep ambient noise to a minimum without degrading intelligibility. Adapting a good-quality audio compressor such as the RP Electronics RPC-3* to the expansion mode is extremely easy as shown in fig. 1.

Thirdly, an important factor in

1. Harold G. Collins, W6JES, "Ordinary and Processed Speech in SSB Application," *QST*, January, 1969, page 17.

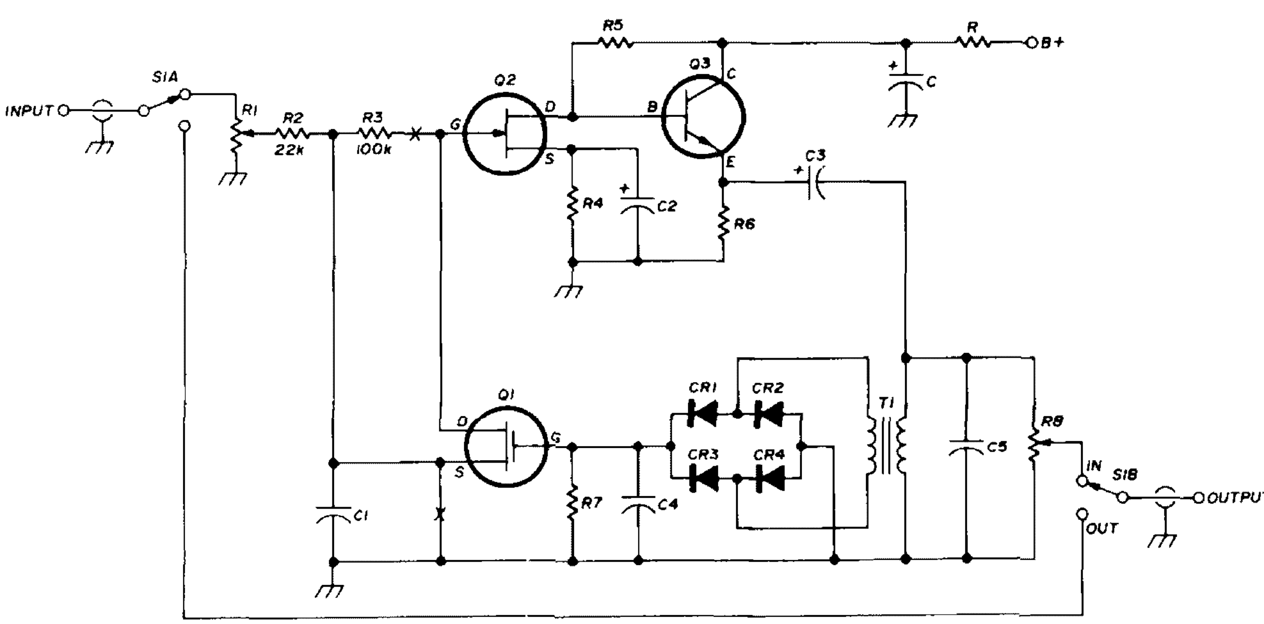


fig. 1. Basic circuit of the RP Electronics RPC-3 speech processor. Expand function is added by breaking connections between R3 and Q2 gate and between Q1 source and ground and connecting Q1 source to the junction of R2 and C1.

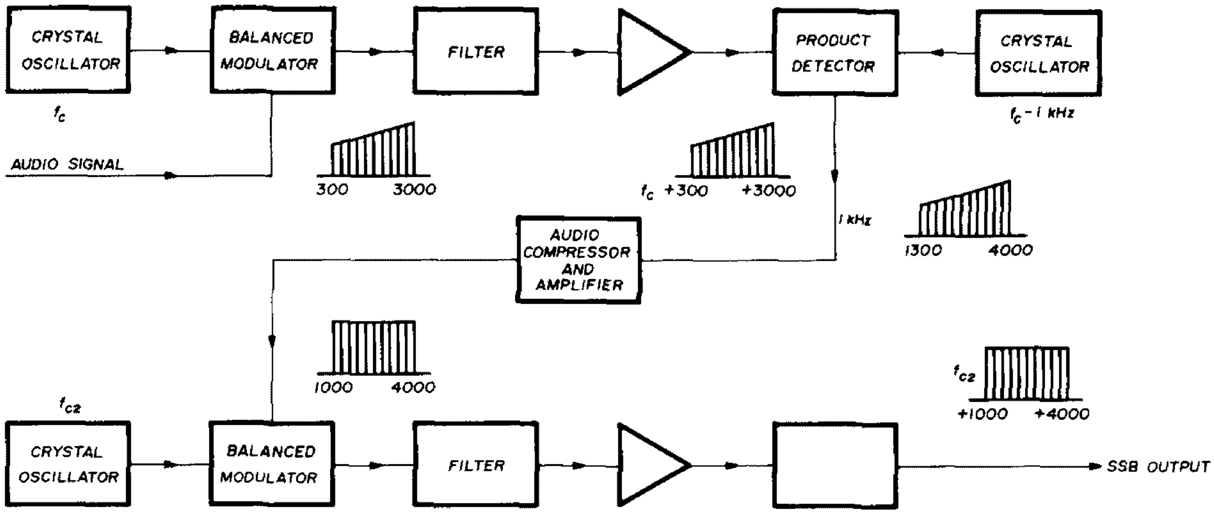


fig. 2. Block diagram of PAØKT's method of achieving an ssb signal with constant amplitude using a fast-acting audio compressor with an offset technique that produces a residual carrier at 1 kHz so that full output can be obtained during pauses in speech.

communications theory (but almost entirely overlooked in the amateur literature) is the role of redundancy in effective transmission of information. On several occasions I chanced to overhear a weak signal of a young amateur who was using a reverbration system in the audio string and was very much impressed at the apparent readability improvement of his signal. I have also heard the use of reverbration by foreign broadcast stations with apparent improvement of intelligibility. Parity testing, a form of redundancy, is standard practice in computer data transmission.

Finally, it has been mentioned that rf clipping simulates a form of variable pulse-width modulation, which is essentially digital, as opposed to the analog waveform characteristics of unprocessed audio. With the recent introduction of relatively low-cost but powerful and fast mini-computer systems, it may be feasible at this time to develop a real-time speech-processing algorithm for precise computerized control of speech processing parameters.

James G. Limber, K9ZAT
Chicago, Illinois

*Available from RP Electronics, Box 1201, Champaign, Illinois 61820.

rf interference

Dear HR:

One common type of RFI to which hi-fi equipment is susceptible is the "thumps and bangs" which come from the thermostats in refrigerators and central heating systems. It is not generally realized that these noises are usually caused by rectification of the radiofrequency component of the unwanted signals, and that any hi-fi equipment which is susceptible to "fridge crunch" is almost certainly also susceptible to other forms of rf interference.

If your neighbor complains of interference, it is a good approach psychologically to say something along the lines of, "Yes, it is a problem with some hi-fi equipment — it probably picks up your refrigerator as well." Nine times out of ten you will hit the nail on the head and get your point across. Your neighbor's displeasure at having to get his hi-fi fixed to remove your transmissions will be reduced if you point out that standard RFI measures will

usually remove other unwanted noises as well.

clipping and rfi

It may come as a surprise that while a speech clipper makes your signal sound louder to other amateurs, it actually reduces the level of interference picked up on high-fidelity installation. With an ssb signal the hi-fi system registers the difference between the peaks and the troughs of your voice, and so theoretically if you clip sufficiently (enough to lift the noise between words to peak level), all amplitude variations will be cancelled, and the RFI will disappear.

In practice, if an rf speech clipper is used with around 6 to 10 dB of high-frequency pre-emphasis,[2] about 30 dB of clipping can be applied without objectionable distortion. At this level of clipping, unless you run a *very substantial* linear, input power has to be reduced somewhat and a combination of this with the clipping can result in a considerable reduction in RFI along with a net gain in talk power.

Some experimenting has been done along these lines in Europe where the clipping has been taken to the extreme of being infinite – a block diagram of the system developed by PAØKT is shown in fig. 2.[3] Although all the audio components are at the same level, the signal is still quite readable. This may seem like a drastic approach, but in difficult interference problems where unsympathetic licensing authorities are involved, it has solved the problem.

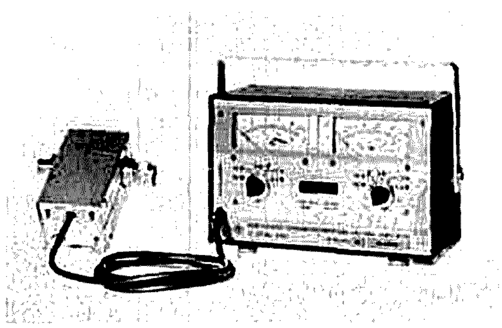
Harry Leeming, G3LLL
Holdings Audio Center
Blackburn, England

2. B. Kirkwood, ZL1BN, "Principles of Speech Processing," *ham radio*, February, 1975, page 28.

3. P. Hawker, G3VA, "Constant-Amplitude SSB," *Radio Communicatios*, November, 1974, page 762.

new products

vhf/uhf directional rf power meter



The new Rohde & Schwarz directional rf power meter, type NAUS-80, provides simultaneous measurement of incident and reflected power over the frequency range from 25 to 1000 MHz without any switching or changing measuring heads. The indication accuracy of the power meter is within 4% of the *reading* and ±1% of full scale – this is a vast improvement over the accuracy of most rf power meters which is usually specified only as a percentage of full-scale deflection. Although the NAUS-80 power meter is designed for operation over the range from 25 to 525 MHz, it is usable to 1000 MHz. If desired, the factory can calibrate the instrument to 1500 MHz at slight additional charge. Power ranges are 3.2, 10, 32, 100 and 320 watts full scale.

The NAUS-80 rf power meter consists of two units: the measuring head, and the indicating unit. The measuring head contains a symmetrical directional coupler which measures both incident and reflected power. Networks within the measuring head compensate for the voltage coupled out, which rises with frequency. The coupling attenuation and voltage division in the directional coupler are adjusted so that the rf rectifying diodes operate only in the square-law region. This permits the use of easy-to-read, linear meter scales.

The small rectified voltage (10 μV to 25 mV) from the directional coupler is amplified in a chopper amplifier which converts the dc input voltage to a square wave which is boosted 50 dB in a series-connected amplifier. The amplified signal is then applied to an attenuator which is ganged with another attenuator in the feedback path. In the 3.2 watt position attenuation is 0 dB, increasing 10 dB with each measurement range. Since the attenuation in the feedback path is reduced simultaneously in corresponding amounts to the main attenuator, the loop gain of the chopper amplifier is the same on all measurement ranges so the meter indications are free of oscillation and transient response remains constant.

The attenuator is followed by the final amplifier where the signal is boosted by another 50 dB and then fed to a synchronous detector. The transistors in the synchronous detector are driven together with the same square-wave generator which drives the chopper amplifier. The synchronous detector operates into a charging capacitor; a series resistor is included to form a lowpass filter with a low cutoff frequency. The voltage at the output of the lowpass filter (approximately 300 mV on all measurement ranges for full-scale deflection) is connected to the panel meter.

The feedback voltage to the chopper amplifier is fed back through a thermistor which compensates for the slight

More Details? CHECK–OFF Page 126

temperature effect on the rf rectifying diode in the directional coupler. Temperature effect on the meter indication is less than 0.25% (referenced to indication at 25°C).

Although the directivity of the directional coupler is 30 dB or more above 30 MHz, finite reflected power would be indicated when working with matched terminations. The designers compensated for this effect by connecting the inverting output of one channel to the non-inverting input of the other channel through resistors. The rectified voltage of the incident power being measured in channel A is attenuated such that the voltage available at the inverting input of the channel B chopper (reverse power channel) is virtually the same as the voltage present at the non-inverting input which develops because of the finite directivity of the directional coupler. The two voltages balance out and, as a result, there is no indication on the reflected power meter. The effect is the same as if the measurement were made with a match-terminated directional coupler with infinite directivity.

Our staff had an opportunity to evaluate the NAUS-80 recently, and it proved invaluable in setting up the tuned input circuit of a high-power two-meter linear for minimum vswr, and measuring drive power. Its high reflected power sensitivity and panel meters for both incident and reflected power are especially helpful in quickly evaluating the effect of circuit adjustments. Since the instrument is completely portable, it can also be taken to the top of your tower for precise antenna checks.

The Rohde & Schwarz NAUS-80 rf power meter is available for 50, 60 or 75 ohms. Various types of connectors are available including type N, BNC and UHF, and may be changed in the field. The built-in power supply uses five 1.5-volt D-size dry cells with an estimated operating life greater than 7000 hours (total power drain is 1 mA or less,

depending upon the measurement range). The vswr of the measuring head is 1.03 or less, and insertion loss up to 300 MHz is 0.1 dB or less (0.25 dB or less to 525 MHz). The instrument is covered by a five-year warranty.

Also available from Rohde and Schwarz is the type NAN high-frequency wattmeter and matching indicator which provides direct power and reflection coefficient measurements over the high-frequency range. Accuracy is within 5% of full scale and frequency response is flat within 3% over the range from 1.5 to 30 MHz. The instrument is available with a characteristic impedance of 50, 60 or 75 ohms. Reflection due to the coupling system is less than 2%. The instrument has four selectable power scales up to a maximum of 1200 watts.

For more information on the NAUS or NAN rf power meters, write to Rohde & Schwartz Sales Company, 14 Gloria Lane, Fairfield, New Jersey 07006 or use *check-off* on page 126.

two-meter fm transceiver

The Horizon 2 is a new, 12-channel, 25-watt two-meter radio developed by Standard Communications Corporation. This rig is an outgrowth of Standard Communications' land/mobile and maritime equipments, which must meet rigid FCC type acceptance requirements for the transmitter section. The receiver section meets the proposed maritime FCC receiver specifications as well as the current receiver DOC type acceptance requirements for use in Canada.

Some of the features of the Horizon 2 include 25 watts *nominal* output; 23 watts *minimum*. The unit is capable of using 12 channels; three are included: 94/94, 52/52, and 16/76. Crystal net capacitors are included for both transmit and receive. The receiver front end has a selective ceramic filter that provides −65 dB minimum selectivity.

More Details? CHECK—OFF Page 126

Plenty of audio power is available – more than 3 watts – perfect for the noisiest mobile installation. The rig is center tuned to 146.94 MHz and will operate on the low and high ends of CAP and MARS frequencies.

The Horizon-2 amateur net price is $225.00, which includes the three channels mentioned above. For other channel options and more data, write Standard Communications Corporation, P.O. Box 92151, Los Angeles, California 90009, or use *check-off* on page 126.

250-MHz frequency counter

The K-Enterprises model 4X6 six-digit frequency counter covers the frequency range from 500 kHz to 250 MHz with sensitivity of 80 mV or less at 150 MHz. The input impedance of the counter is 50 ohms and maximum input voltage is 15 Vrms or 50 Vdc. The time base uses a crystal clock with an accuracy of 10 ppm over the temperature range from zero to +40°C. The model 4X6, which contains a built-in power supply for operation from 117 Vac, is priced at $250. The model 4X6C, which includes a temperature compensated crystal oscillator with 0.0005% accuracy from −30° to +60°C, is priced at $270. Add $2.50 to cover postage and insurance.

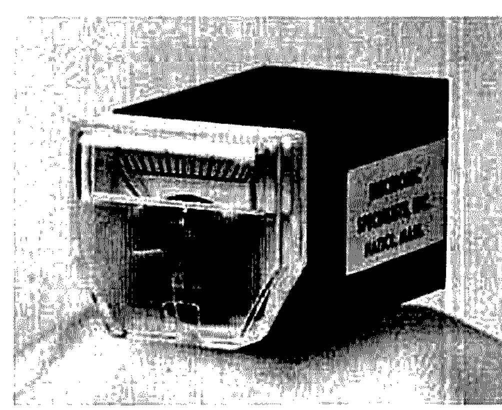
For more information on the K-Enterprises 250-MHz frequency counters, write to them at 1401 East Highland, Shawnee, Oklahoma 74801, or use *check-off* on page 126.

fm power amplifier

Specialty Communications Systems has introduced a new Porta-Pack 25-watt power amplifier system with a self-contained battery pack which can be carried over your shoulder. Designed for use with the popular H-T fm trans-

ceiver, the amplifier can also be used with other similar hand-held fm transceivers, and covers the complete two-meter band without retuning. The battery pack provides 45 minutes of "talk" time and can be recharged in four hours. Included with the $199.95 package price are the 25-watt power amplifier, 12-volt Gates cell energy source, external constant-voltage charger, flexible antenna and a leather case with belt loop and shoulder strap. For more information, write to Louis Anciaux, WB6NMT, Specialty Communications Systems, 4519 Narragansett Avenue, San Diego, California 92107, or use *check-off* on page 126.

radio-sentry mini-meter



Mini-Meter, a new amateur fm transmitter monitor, has been introduced by Electronic Specialists. The transmitter field strength is continuously monitored, with the status displayed on a miniature meter. Ultra compact, *Mini-Meter* operates without batteries or wires and can be carried in your pocket for on-the-spot transmitter checks. Deteriorating performance can be spotted early, allowing timely maintenance. The unit is supplied with a convenient, detachable mounting arrangement. State frequency. $27.95 postpaid from Electronic Specialists, Box 122, Natick, Massachusetts 01760.

More Details? CHECK–OFF Page 126

$1.00

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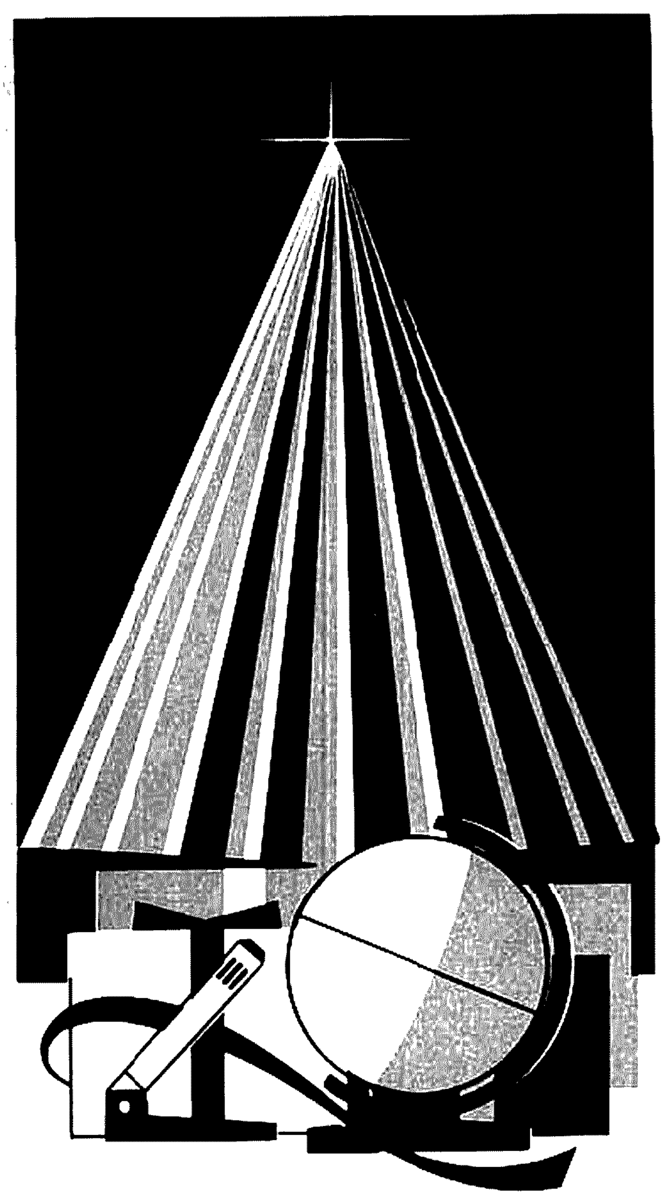

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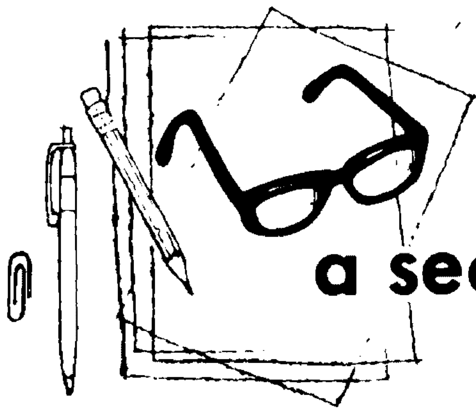


a second look

by jim fisk

Beginning this month we are presenting a series of articles on microprocessors by Dave Larsen, WB4HYJ, Peter Rony and Jonathan Titus, authors of the popular series of *Bugbooks.** Not since the development of the transistor in 1948 has any product or technology offered such an exciting promise of things to come as the microprocessor — literally a computer on a chip.

Computers in the 1960s are credited with revolutionizing the engineering and accounting fields by replacing people power with instantaneous electronic computation and retrieval. Microcomputers in the 1970s are expected to extend these benefits into areas where existing computer technology has never before penetrated, including amateur radio. Several groups are now working on microprocessor controlled vhf-fm repeaters, future OSCAR satellites will carry an on-board microprocessor for systems maintenance and control, and VE3SAT and others are already using microprocessors for ASCII communications through OSCARs 6 and 7.

Other amateur applications such as RTTY speed control, RTTY-ASCII or RTTY-Morse conversion, and automatic Morse code copiers are a natural for microprocessors. Automatic satellite tracking systems, log keeping, transmitter tuneup and control, and antenna pointing systems are other straight forward microprocessor-based systems which will see widespread use in the future. If, for example, you're a DXer and hear a rare VP8 on 20 meters, you would just punch VP8 into your keyboard and your beam would automatically come around to the correct heading. If you were operating on CW you would only have to tap out VP8 in Morse code — the microprocessor would convert the Morse characters into machine language, translate that into a beam heading, and turn on your antenna rotator.

Until recently the cost of microprocessor chips put them out of reach for most amateur applications, but as more and more manufacturers have gotten into the act the prices have dropped dramatically. The popular 8-bit 8080 microprocessor which was originally developed by Intel, for example, was selling for $300 to $400 a little more than a year ago, dropped to about $150 this past summer, and is now available from one source for under $30. Although these prices are still a bit high for the amateur experimenter, industry sources predict that microprocessors will sell for $5 or less within a couple of years, perhaps as early as 1977.

In addition to the microprocessor series in the magazine which is designed to familiarize amateurs with this important new technology, during 1976 *ham radio* will be presenting a series of one-day microprocessor seminars at various hamfests across the country including SAROC in Las Vegas (January 9th), Miami (January 24th) and Dayton (April 23rd and 24th). The fee for the seminar is $50 and includes $35 worth of books. Since seating is limited, early registration is recommended — write to *ham radio* for details.

Jim Fisk, W1DTY
editor-in-chief

**Bugbook I and II, Logic and Memory Circuits Using TTL Integrated Circuits; Bugbook III, Microcomputer Interfacing Experiments using the Mark 80, an 8080 system*, $35 the set from Ham Radio Books, Greenville, New Hampshire, 03048.

presstop

ANY HOPE THAT DOCKET 20282 — Restructuring — would be out by the end of this year has been much too optimistic. FCC Safety and Special Services Chief, Charles Higginbotham, W3CAH, feels that sometime next spring is a much more realistic target, and even then some aspects may require reexamination as additional dockets or oral proceedings such as the ARRL has requested. A tremendous amount of work has already gone into analyzing the mountain of Comments with the task far from done, and problems associated with CB's explosion aren't helping the effort.

SECRECY PROVISIONS of the Communications Act of 1934, Section 605, deserve a lot more attention by Amateurs than they've been getting. A strict interpretation of Section 605 forbids the disclosure of anything heard on the air except broadcast and Amateur transmissions — and that includes mentions of frequencies or any other information regarding the overheard signals!

Since This Ban applies to CB as well as other services, it could put a severe crimp in some of the recently publicized CB clean-up efforts conducted by Amateur groups.

The Intent Of Section 605 is quite clear — how likely an Amateur is to be cited for violating it is not.

REQUIREMENT FOR MULTIPLE COPIES for submissions to the FCC was upheld by Commissioners after consideration of a petition for its elimination submitted by W6NJU. Additional copies are necessary to insure the submission reaches all who should see it, but in their review of the requirement reductions were found possible.

Effective October 14 the number of copies required for comments on a Notice of Proposed Rule Making was reduced from 15 to 12 (original plus 11 copies) — other requirements not likely to affect Amateurs were also reduced. In their rejection of W6NJU's petition the Commissioners also noted that single copy submissions are now and have been accepted although they do not receive as wide circulation as those that meet the requirement.

Ham Radio/HR Report readers should not forget our long standing public service offer. Send your FCC submission directly to us and we'll make all the necessary copies and mail them to the Commission for just $1.00 per page of original document.

REPEATER FUNDING may become an issue with the FCC if some flagrant abuses aren't corrected. Though use of a club's dues to pay for repeater maintenance is well within the Amateur rules, the solicitation of money for the use of a given repeater or its facilities (such as autopatch) is almost certainly a violation of Part 97.112, "No remuneration for use of station."

OSCAR ORBITAL PREDICTION BOOKLET produced by W6PAJ will replace HR Report's monthly prediction sheets for HR Report subscribers in 1976. W6PAJ's handy booklet will be sent without charge to any subscriber who asks for it — dropping the monthly sheets was done in recognition that a vast number of subscribers did not use them and HR Report sheets were a duplication of effort.

6000 MILE OSCAR QSO was completed between G3IOR and W6CG! Using meteor scatter techniques on selected orbits as the Satellite was over the horizon between them, successful two-way communications were finally exchanged between the two over a period of two weeks. Congratulations to both!

CIVIL SERVICE ADMINISTERED Amateur Radio exams have not been as popular in the test areas as expected. At the mid-point of the two-year program (which runs until next July) no specific conclusions have been drawn and FCC Field Operations people are watching it carefully.

BARRY ELECTRONICS WILL CONTINUE as a major Amateur Radio supplier despite Barry's tragic loss in a boating accident on Long Island Sound. Barry's wife Kitty vows she and the crew will keep the business going just as before.

LA1SH/BY QSLs received by several west coasters are pretty exciting wallpaper but little else. It's now considered certain that he operated only from shipboard.

frequency synthesizer

for the Collins 75S receiver

Complete description of a frequency synthesizer that converts the Collins 75S-series of communications receivers to general-coverage use from 3.4 to 30 MHz

Robert S. Stein, W6NBI*

During the past few years, the use and application of frequency synthesis in both receivers and transmitters has increased tremendously. A quick perusal of the ads for vhf transceivers is all the evidence needed to verify this fact, although there are also many high-frequency military and commercial (non-amateur) transmitters and receivers which employ frequency synthesizers to generate specific frequencies required within those units.

The advantages of having a general-coverage receiver in the ham shack are manifold and were discussed in a previous article describing a synthesizer for use with the Drake R-4 series receivers[1]. This article will describe a frequency synthesizer to supplement or replace the high-frequency oscillator crystals in a Collins 75S-1, 75S-2 or 75S-3, resulting in a receiver which covers 3.4 through

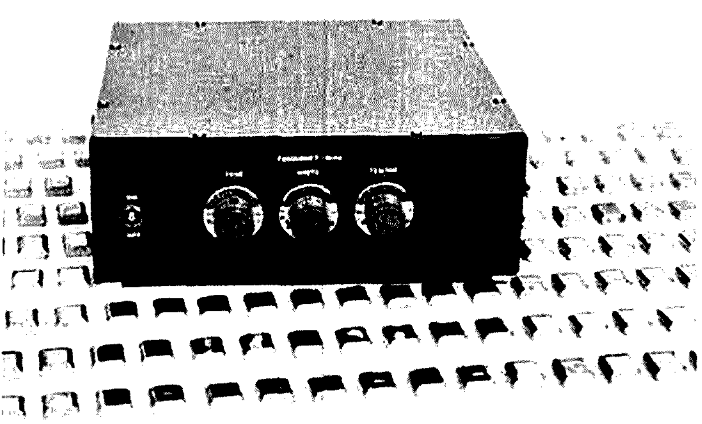
*1849 Middleton Avenue, Los Altos, California 94022.

30 MHz. Only minor electrical changes are required in the receiver; no holes need be drilled nor other mechanical modifications made.

Before proceeding with the description of the frequency synthesizer which makes this possible, a review of the receiver conversion process is in order. The 75S-1, 75S-2 and 75S-3 receivers all utilize identical crystal-oscillator and first-mixer circuits, so that the discussion is applicable to any one of the receivers.

The local-oscillator (LO) frequency injected into the first mixer is 3.155 MHz higher than the low-frequency end of the desired 200-kHz tuning range. Since the high-frequency oscillator is crystal controlled, this requirement is translated to a crystal which will generate the proper frequency. For example, the 3.8- to 4.0-MHz band requires a 6.955-MHz crystal (3.8 plus 3.155 MHz), which is supplied with the receiver. Collins specifies that the receiver

S-line frequency synthesizer with some of the 128 crystals from the Collins CP-1 crystal pack, which it replaces.



will tune from 3.4 to 30 MHz with the proper crystal. It should be noted that for receiver frequencies above 12 MHz, frequency doubling takes place in the plate circuit of the crystal oscillator; therefore the *crystal* frequency is one-half the frequency injected into the first mixer. Nevertheless, the *injection* frequency is always 3.155 MHz above the lower band edge.

In order to cover the entire range of 3.4 to 30 MHz in 200-kHz increments, 133 crystals would be required, starting with a 6.555-MHz crystal for the 3.4- to 3.6-MHz band, a 6.755 MHz crystal for the 3.6- to 3.8-MHz band, and so on. Even with the 28 crystal positions available in the 75S-3A, it is obvious that a complete set of crystals would not only be impractical to use, but prohibitively expensive. However, if we can generate frequencies of 6.155 through 32.955 MHz every 200 kHz, and substitute them for the crystals in the hf crystal oscillator, we can achieve all-band coverage within the specified tuning range of the receiver. The frequency synthesizer to be described does exactly that.

basic phase-locked loop frequency synthesizer

Although the basic phase-locked loop frequency synthesizer has been explained in previous articles, a brief review at this time will simplify the detailed explanation of this specific synthesizer. Fig. 1 shows the basic phase-locked frequency synthesizer. A stable reference frequency is applied to one input of a phase comparator. The output of the phase comparator is a dc voltage which passes through a lowpass filter and controls the frequency of a voltage-controlled oscillator (vco). The oscillator generates the desired frequency, which may be any multiple of

the reference frequency. The vco output is also applied to a frequency divider whose function is to divide the vco output frequency to the same frequency as that of the reference oscillator.

Let's assume that the reference oscillator frequency is exactly 5 kHz and that an output frequency of 6555 kHz is required. If we have a divider or programmable counter which will divide by 1311, the signal input to the phase comparator will also be 5 kHz when the vco output is exactly 6555 kHz. This is accomplished by the phase comparator producing a dc output which "tunes" the vco until it is exactly 6555 kHz. The divided vco frequency is then exactly 5 kHz, the same as the reference frequency. Thereafter, the vco output will stay at 6555 kHz; any variation from this frequency changes the signal input to the phase comparator, which in turn produces a dc output change and brings the vco back to 6555 kHz. Thus, the output frequency is locked to the reference frequency, and has essentially the same stability as the reference oscillator.

By using a frequency divider which can be programmed, it is possible to obtain virtually any number of discrete frequencies which are integral multiples of the reference frequency, all of which are phase locked to the reference oscillator. The lowpass filter keeps the reference frequency from modulating the vco and establishes the lock-up time of the loop.

75S synthesizer

A block diagram of the 75S hfo frequency synthesizer is shown in fig. 2. The loop reference is a 100-kHz crystal-controlled oscillator, which is divided by ten and then by two, resulting in a 5-kHz reference signal which is applied to one input of the phase comparator. The other input to the comparator is the divided vco frequency, which will be discussed presently. The output from the comparator is a function of the difference between the two input frequencies and is applied to the loop filter, consisting of an active and a passive lowpass filter. The resultant dc controls the vco frequency by changing the capacitance of a varactor diode. The vco output, 6.555 to 32.955 MHz in 200-kHz steps, is amplified to a suitable level and routed to the receiver.

The vco output is also applied, via an isolating source follower, to a Schmitt trigger, which converts the amplitude and waveform of the vco output to one that is compatible with the TTL integrated circuits in the frequency divider.

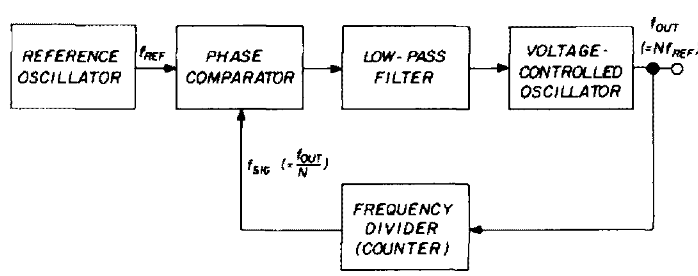


fig. 1. Basic phase-locked frequency synthesizer. The frequency divider is a variable-modulus, or programmable, counter.

The vco frequency divider is a variable-modulus counter which can be programmed to divide by any factor between 1311 and 6591 in steps of 40 (i.e. 1311, 1351, 1391, 1431, etc.). An examination of the discrete vco frequencies to be synthesized will reveal that the largest common factor is 5 kHz, thereby establishing the reference frequency. Steps of 200 kHz in the vco output are obtained by changing the counter modulus in steps of 40 (40 x 5 kHz = 200 kHz). Since each vco frequency ends in 5, the least significant digit in the number by which the vco frequency must be divided to yield 5 kHz will always be 1. Therefore the first counter always provides a 1-count. The three remaining counters are programmed by the front-panel frequency-

control switches and establish the first three digits of the frequency divisor.

Preselected binary-coded-decimal (BCD) outputs from each counter, plus the output from the Schmitt trigger, are fed to a decoder circuit, which produces Trimmer capacitor C3, in series with the crystal, permits adjustment of the crystal frequency. The output of the oscillator is shaped and buffered by a third gate, U1C.

The 100-kHz signal is divided down

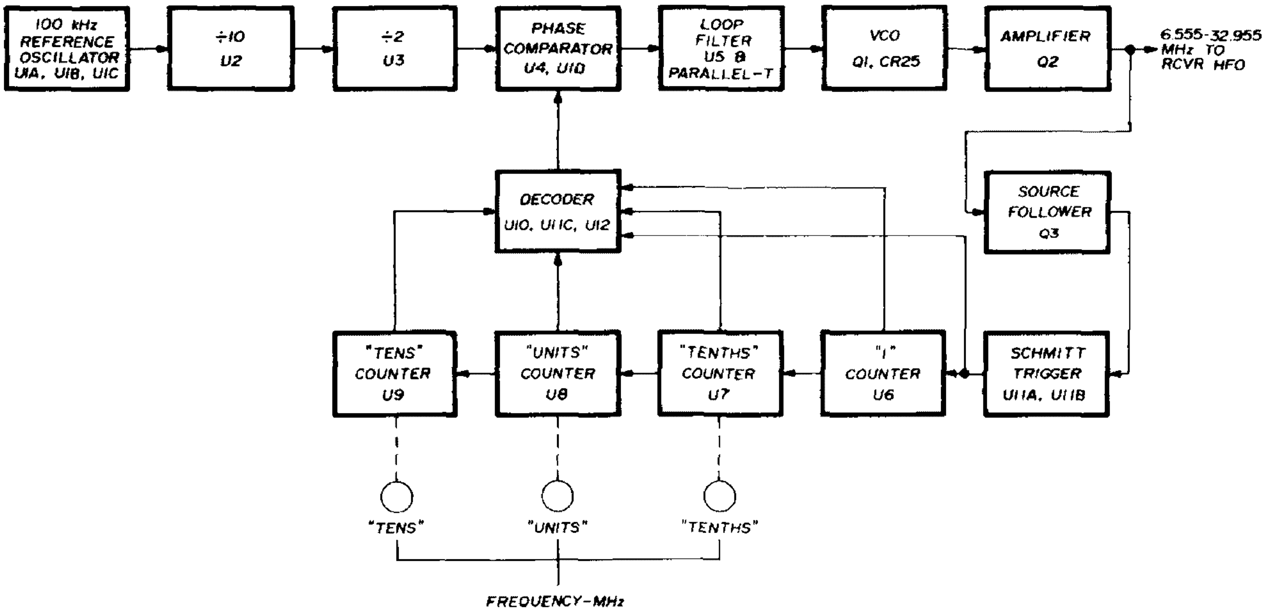


fig. 2. Block diagram of the Collins 75S hfo frequency synthesizer. The only tuning controls are the front-panel TENS, UNITS, and TENTHS rotary switches.

the divided-down vco signal applied to the phase comparator. The decoder also resets the counters (actually this is its primary function), but that signal path has been omitted from fig. 2 because it is not pertinent to overall signal flow. Details of the counter and decoder functions will be explained in greater detail under their circuit descriptions.

reference oscillator and phase comparator

The 100-kHz reference oscillator and its frequency dividers, the phase comparator, and the loop filter are shown in fig. 3. The reference oscillator consists of gates U1A and U1B, two sections of an MC846P quad 2-input NAND gate. The gates are configured as a multivibrator and the 100-kHz signal is developed by connecting crystal Y1 as part of the signal path between the gates. to 10 kHz by U2, a 7490 decade counter. The output of the 7490 is then divided by two by one flip-flop in U3, a 7474 dual D-type flip-flop, resulting in the 5-kHz reference which is applied to pin 3 of phase comparator U4.

The phase comparator comprises U4, another 7474 dual D-type flip-flop, and the remaining gate section of U1. It compares the phase difference between the 5-kHz reference and the vco frequency divided by the counter modulus (f_{vco}/N), and produces a digital pulse output whose duty cycle is a function of the phase difference. This digital output is partially filtered by R3 and C6 to a sawtooth which is applied to the inverting input of U5.

U5 is an LM3900 quad op amp, one amplifier section of which is used as the active element in the loop filter. It attenuates the ac components of the sig-

nal from the phase comparator and thereby produces a dc output which varies in accordance with the phase difference between the inputs to the com-will be improved.) Additional attenuation of harmonics of the reference frequency is accomplished by R22, R34, C18 and C38 in the vco (fig. 4).

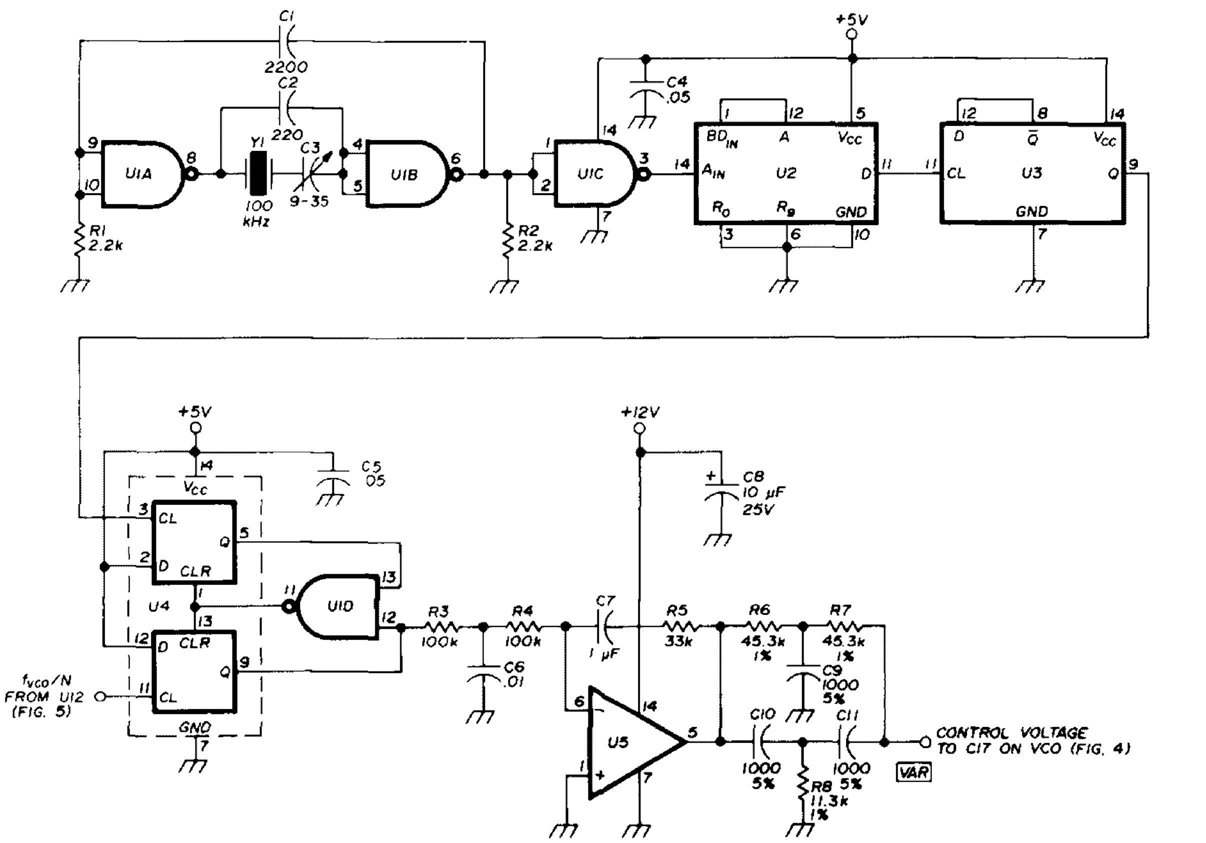


fig. 3. Schematic of the crystal oscillator and its frequency dividers, the phase comparator, and the loop filter. Integrated circuits are listed in table 1 (page 20). C7 must be a polycarbonate- or polyester-film type capacitor.

parator. The gain of the loop filter, its frequency response, and the loop lock-up time are determined by the values of R3, R4, R5 and C7[2,3].

Additional filtering of the 5-kHz loop reference frequency is needed to prevent modulation of the vco, which would produce spurious sidebands on both sides of the desired frequency. A parallel-T filter, consisting of R6 through R8 and C9 through C11, provides a minimum of 35 dB attenuation at 5 kHz. (This figure is based on worst-case conditions using five-percent capacitors. If two- or one-percent capacitors are used, or the capacitors are selected by bridge measurement, the attenuation

voltage-controlled oscillator

The vco is built as a separate, shielded unit to eliminate stray pick-up from the digital circuits and from ac fields. The oscillator consists of Q1, an E300 (or equivalent) n-channel fet, in a Colpitts circuit with varactor CR25 connected in series with C38 across the tank circuit. The varactor is a Motorola MV1401 and has a ratio of maximum-to-minimum capacitance of approximately ten, as compared to usual ratios of two to four for conventional varactors. (It also happens to be the most expensive single component in the entire synthesizer.) Despite the large capacitance ratio, the oscillator cannot

cover the entire range of 6.555 to 32.955 MHz without switching. This is accomplished by diode switching, using switch section S3-D of the *tens* divider switch (fig. 5).

In the zero position of the *tens* switch, diodes CR26 and CR27 do not conduct, so coils L1 and L2 are each effectively in series with a 33-μH choke (L4 and L5) to ground. The high value of this inductance has only stray effect on the circuit; thus the oscillator frequency is essentially determined by coil L3 and the tank-circuit capacitance. When the *tens* switch is set to position 1 or 2, one of the diodes is biased into forward conduction and brings the low end of the associated coil close to rf ground, shunting L3 and thereby lowering the tank-circuit inductance. Resistor R21 in series with the switch arm limits diode current to a safe value.

The oscillator output is taken from the source of Q1 and coupled to the base of amplifier Q2. The amplifier, a type 2N2219 npn transistor, is a broadband stage which feeds the hf oscillator circuit in the receiver through an isolating 5-dB L-pad, R27 and R28. Also applied to the output circuit is the +12-volt power supply, which is decoupled from the vco signal by rf choke L8 in series with current-limiting resistor R32. This dc source is used to actuate a sensitive relay in the receiver, as will be explained later.

The output of Q2 is also coupled to Q3, an n-channel fet configured as a source follower. The source load is made up of two resistors, R30 and R31, which form a 6-dB L-pad in the output. The source follower drives the Schmitt trigger in the digital portion of the synthesizer and, with the L-pad, keeps

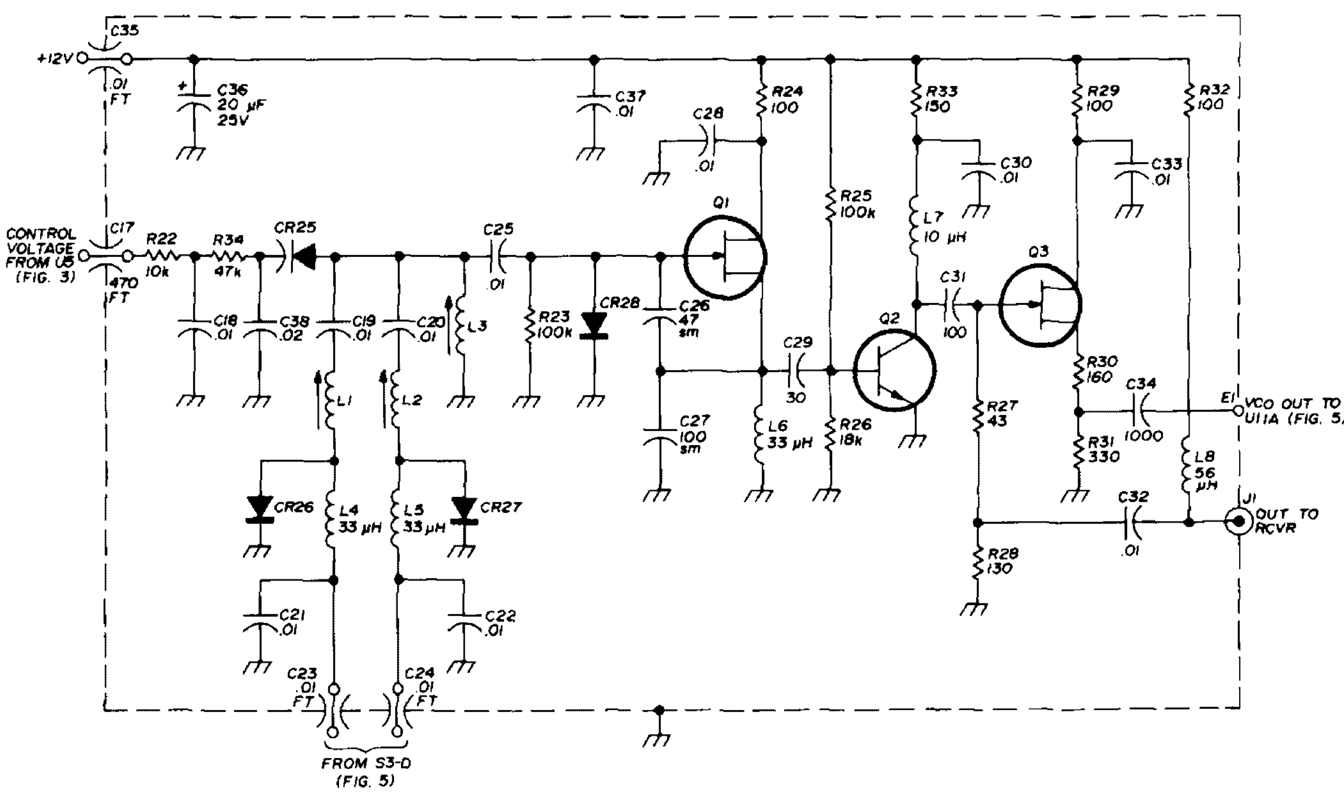


fig. 4. Schematic of the vco. See table 1 for coil-winding data and descriptions of parts not identified on the schematic.

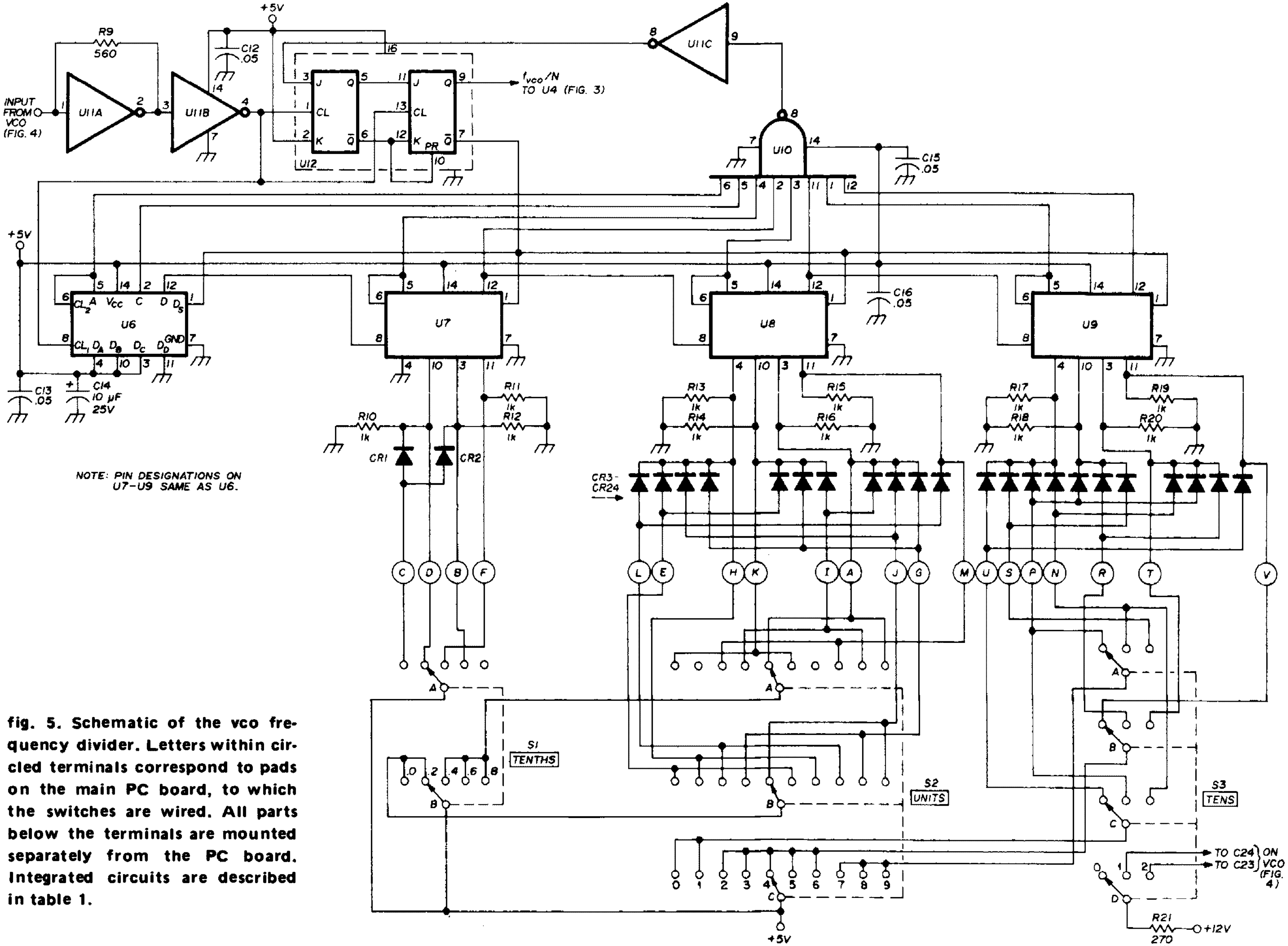


fig. 5. Schematic of the vco frequency divider. Letters within circled terminals correspond to pads on the main PC board, to which the switches are wired. All parts below the terminals are mounted separately from the PC board. Integrated circuits are described in table 1.

any digital signals from feeding back into the vco.

frequency divider

Fig. 5 shows the vco frequency divider and its associated front-panel switches. Note the use of the word *divider* in its singular form; the counters used in the divider circuit function as an integral circuit (no pun intended), rather than as separate divider stages such as are used to divide the 100-kHz crystal frequency down to 5 kHz. Because this is quite different from the usual frequency multiplier or divider stages familiar to most amateurs, as evidenced by the many inquiries received following publication of the R-4 synthesizer article[1], it seems appropriate at this point to explain the operation of a typical variable-modulus, or programmable, counter.

Let us consider a basic two-stage frequency divider, as shown in **fig. 6**. Each of the counters is a decade counter, that is, a counter which produces one output pulse for every ten input clock pulses. However, each counter is presettable, which means that its count may be programmed or modified by setting its data inputs (D_A, D_B D_C, and D_D) either high or low. The data-input subscripts indicate the binary weighting assigned to each input: A=1, B=2, C=4, and D=8. There is also a fifth data terminal, D_S; this is the data-enable input, which must be set *prior to and during* the interval that the data inputs are applied. In the simplified circuit shown, we will assume that D_S must be set high to enable the data inputs.

Conventional digital terminology designates the first pulse in a pulse train as 0, so that the tenth pulse, which produces an output from a decade counter, is therefore designated number 9. The total number of clock pulses, N_{max}, which can be counted before an output is produced from a ripple-through counter (another name for the circuit shown in **fig. 6**) is

$$N_{max} = N_1 \times N_2 \times \ldots N_n$$

where N_1 is the modulus of the first counter, N_2 is the modulus of the second counter, and so on. Since each counter in **fig. 6** has a modulus of 10, N_{max} = 100. But remember that this will be clock pulse 99, since we start with pulse 0.

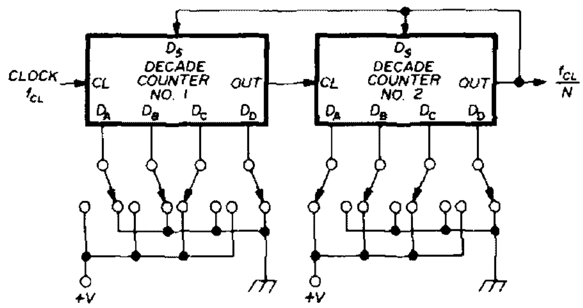


fig. 6. Basic variable-modulus counter.

Now let's assume that we want the circuit of **fig. 6** to divide the clock frequency by 25. If the counters are up-counters (as are all those used in this synthesizer), their data inputs must be preset with the nines' complement of the desired divisor. (Nines' complement simply means the difference between nine and the desired count.) The preset data to be entered, N_D, is determined by the equation

$$N_D = (N_{max} - 1) - D$$

where D is the frequency divisor. Since N_{max} is 100 for our circuit,

$$N_D = (100 - 1) - 25 = 74.$$

The least significant, or units, digit corresponds to the count of the first counter, since it is counting unit clock pulses; the most significant, or tens, digit corresponds to the count of the second counter because it is counting tens

of clock pulses. Therefore counter number 1 must be preset with a 4, and counter number 2 must be preset with a 7. To do this, D_C (having a binary weight of 4) of counter number 1 is set high and all other data inputs are set low. On counter number 2, D_A, D_B and D_C are set high (binary weighting: 1 + 2 + 4 = 7) and D_D is set low. What we have done is to preset the counters so that each is in the state which would exist following the clock pulse having the same number as the preset data. Since counter number 1 has been preset with a 4, it will produce an output after five clock pulses have occurred (corresponding to the 5 in the desired divisor of 25). *Thereafter*, the first counter will count by *ten* until D_S is set high by the output of counter number 2. Similarly, because counter number 2 has been preset with a 7, it will produce an output after it has counted two pulses from the first counter, completing the count of 25. This output is applied to the D_S inputs of both counters and re-enables the data inputs, starting the count over.

If we analyze the operation of the counters, we can see that by presetting the first counter with a 4, the elapsed time between clock pulse 0 and clock pulse 9 was shortened by four clock-pulse intervals. In the same way, by presetting counter number 2 with a 7, the elapsed time between clock pulse 9 (the first output pulse from the first counter) and clock pulse 99 (N_{max}) was shortened by 7 times 10 clock-pulse intervals.. Assuming a clock frequency f_c, with a period t_c,

$$t_{out} = 99t_c - 4t_c - 7(10t_c) = 25t_c$$

and

$$f_{out} = \frac{1}{t_{out}} = \frac{1}{25t_c} = \frac{f_c}{25}$$

The preceding analysis may be extended to any number of cascaded counters and to hexadecimal as well as decade counters. However, actual operation will be limited by the propagation delays through the counters and the set-up times required for the data inputs. As previously stated, the D_S inputs must be enabled *before* and during the time period that the preset data are entered. Since the preset data are dc levels, it follows conversely that they are entered shortly after the generation of the output pulse, which is applied to the D_S inputs. If the clock frequency is too high, the counters may toggle but too much time may elapse, because of propagation delays, between the output pulse following the terminal clock pulse (equivalent to pulse 99 in our basic circuit) and the arrival of the next clock pulse (pulse 0). This will prevent the data inputs from being enabled prior to the arrival of clock pulse 0, and will result in an erroneous count.

Another problem which often arises when using a circuit similar to that of **fig. 6** is caused by the short duration of the output pulse. The output pulse from counters which are used in these circuits has the same width as the clock pulse. Thus the output pulse of counter number 1 is the same as the clock pulse (although with greater time intervals between pulses), and since counter number 2 is toggled by the output of counter number 1, its output pulse width will also be the same as the clock-pulse width. Furthermore, as soon as the output pulse resets the data-enable inputs, both counters resume their preset state and the output pulse disappears. This condition, along with the narrow pulse width, may not permit the data inputs to be enabled for the minimum time which is required by the counter.

The propagation delay may be minimized by decoding the BCD outputs of the counters. These outputs have the same binary weighting as the corresponding data inputs. Thus when the terminal condition of 99 is reached in **fig. 7**, outputs A and D of each counter

will go high, causing the output of the AND gate to go high and enable the data inputs. The advantage of this circuit lies in the reduction in the delay time between clock pulse 99 and the D_S enabling pulse. In fig. 6, the delay is equal to the propagation delay through the two counters, or through a total of eight flip-flops. In **fig.** 7, the delay is equal to the propagation delay through only one flip-flop (flip-flop A in counter number 1) plus that of the gate. This occurs because at clock pulse 98, outputs A and D of counter number 2 are high, as is the D output of counter number 1. Clock pulse 99 needs to propagate only through the first flip-flop in counter number 1 to cause output A to go high, resulting in the required enabling output from the gate.

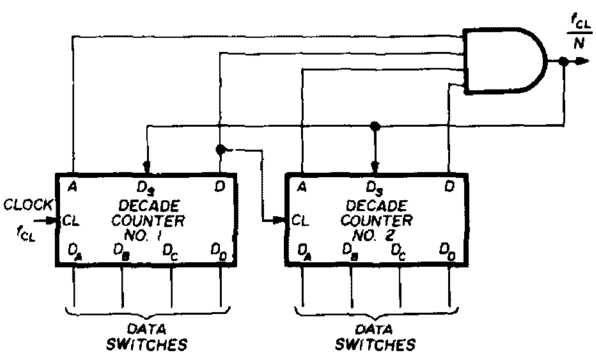


fig. 7. Basic variable-modulus counter with decoded outputs.

This technique of decoding may be used in various circuit configurations. In many cases, it may be necessary to decode outputs which are other than those of the terminal count in order to enable the data inputs before the arrival of clock pulse 0. It may also be necessary to utilize flip-flops in addition to the decoding gate in order to enable the D_S inputs with a pulse whose width is greater than the clock-pulse width. This may result in some preset factors being unusuable, but rarely are all moduli of a variable-modulus counter utilized.

Returning now to **fig.** 5, we see that the vco frequency divider is a four-stage variable-modulus counter comprising U6, a 74S196 or 82S90 presettable decade counter, and U7 through U9, each a 74196 or 8290 presettable decade counter. The signal from source follower Q3 in the vco is converted to TTL level by U11A, one section of a 7404 hex inverter. The inverter functions as a Schmitt trigger by virtue of the connection of R9 between the input and output. Additional shaping is provided by U11B, and the resultant output clocks counter U6 and both flip-flops in U12.

As stated previously, only the counts of the last three counters in the chain need be varied, since U6 provides a fixed count. The counts are controlled by *tenths* switch S1, *units* switch S2, and *tens* switch S3. **Fig.** 5 shows the switches set for a receiver frequency of 4.2 MHz; the corresponding vco frequency is therefore 7.355 MHz. Since the reference frequency is 5 kHz, the vco frequency must be divided by 1471. U6 provides the least significant count of 1. The next significant count of 7 results from setting the data inputs of U7 to the nines' complement of 7, or 2. It can be seen that +5 volts are applied to pin 10 (input D_B) through S1-A, while the remaining data inputs are either grounded directly or are pulled low by resistors R11 and R12. Similarly, it can be seen that U8 provides a count of 4, and U9 a count of 1 for the most significant digit.

The complexity of the switching circuits is the result of labelling the switches so that they indicate the low end of the receiver's 200-kHz tuning range, rather than the dividing count or the vco frequency. Steering diodes CR1 through CR24, in conjunction with the switches, route the 5-volt supply to the appropriate data inputs.

The BCD outputs of the counters are decoded in U10, a 7430 8-input NAND gate, the output of which is inverted by

U11C and applied to pin 3 of U12, a 74S112 dual J-K flip-flop. Both flip-flops are used to re-enable the counters and require three clock pulses after pin 3 is set high by U11C. This means that instead of decoding when the BCD outputs of the counters total 9999, decoding should take place when the BCD outputs total 9996, so that the data inputs of the counters are enabled three clock pulses later, immediately following the terminal BCD state of 9999. However, I found it necessary to shorten the propagation time between the terminal clock pulse and the last-occurring input to U10, and therefore chose to decode a count of 9995. (Note that the A and C outputs from U6 provide the least significant digit of 5.) Since this re-enables the data inputs one clock pulse early, the extra clock pulse is accounted for by presetting U6 for a count of 2, which keeps the least significant digit in the frequency divisor at 1.

Because the preset enabling pulse for the counters must be low, the $\overline{Q}$ output from pin 7 of the second flip-flop in U12 is used. The complementary Q output from pin 9 is applied to the phase comparator for comparison with the reference frequency.

power supply

The synthesizer draws approximately 425 mA from its power supply, which is shown in **fig. 8** along with the interconnections of the main pc board and the vco unit.

Full-wave bridge rectifier CR29 through CR32 is supplied from T1, a 16-volt, 0.5-ampere transformer. The 12-volt supply for op amp U5 on the main board and for the vco is obtained from the output of U13, a fixed 12-volt regulator. The drop to regulated 5 volts for the logic circuitry takes place in U14, a similar 5-volt regulator. LED CR33, connected to the 5-volt supply through current-limiting resistor R35, serves as a pilot light.

construction

Most of the parts comprising the synthesizer are mounted on two printed-circuit boards, the main board and the vco board. The large main board contains all of the digital circuits and the phase detector, while the entire vco, except for the feedthroughs and output connector, is built on the separate vco board. **Figs. 9** through **12** show the foil patterns and parts locations for the two boards*.

The interior photograph shows the construction of the prototype unit, which is enclosed in a *steel* utility box measuring 10x10x3-1/2 inches (25.4x25.4x8.9 cm). It is imperative that a steel enclosure be used because of the strong magnetic field around the receiver caused by the power transformer. The phase-detector circuit and the vco in the synthesizer are extremely sensitive to ac fields, and when the unit was first enclosed in an aluminum box, it was impossible to eliminate a 60-Hz hum on received signals except when the synthesizer was placed directly in front of the receiver. Since this is hardly consistent with good "human engineering" practice, the aluminum housing was discarded and a steel box was substituted.

An enclosure was intentionally selected which was larger than might be expected, because of the sensitivity of the synthesizer to stray fields. This proved to be a wise choice (sometimes you luck out!), since the physical placement of the power transformer within the box became the next problem. Rather than fool around with compartment shielding, I mounted the transformer, rectifiers, and filter capacitor on a small piece of sheet steel. The steel shields the rest of the unit from the transformer

*A set of two drilled and plated boards is available from the author for $14.50, postpaid in the U.S.A. Questions will be answered if accompanied by a self-addressed, stamped envelope.

field when the power supply assembly is positioned so that the steel plate is vertical and the transformer is on the side away from the main board.

Even with the arrangement described, the placement of the power supply assembly is critical. I suggest that the assembly be mounted temporarily, left-hand wall of the enclosure, with their pins extending into the interior. Apply a thin layer of silicone heat-transfer compound between the regulators and the housing to aid in dissipating heat.

The three rotary switches, the power switch, and the LED pilot light are

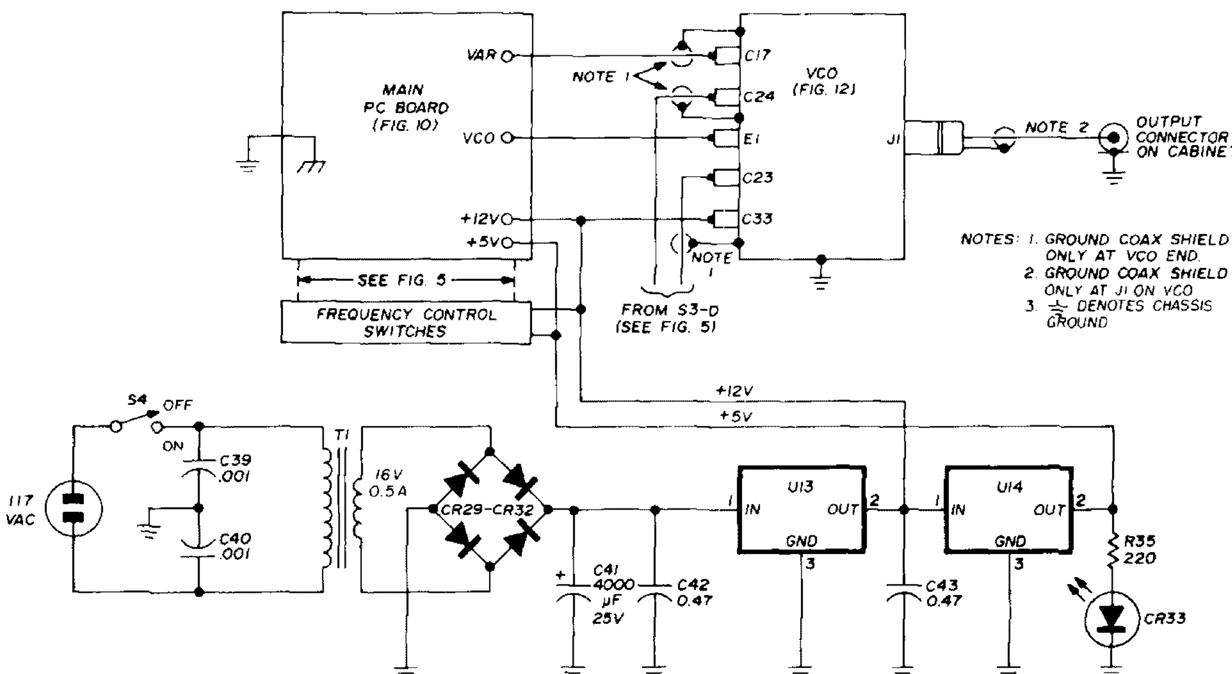


fig. 8. Schematic of the power supply and interconnections of the assemblies comprising the synthesizer. Diodes and integrated circuits are described in table 1.

with long leads, when the unit is constructed. Then after all adjustments have been made and the synthesizer is functioning, the final location can be determined by moving the assembly around until any hum on a received signal disappears. You may find that moving it one way or the other by as little as a half-inch (1 cm) may make a big difference.

On the other hand, this problem can be eliminated, and a smaller cabinet used, if the power supply is made a separate unit and connected to the synthesizer proper by means of a cable. The choice is yours.

The two voltage regulators (U13 and U14) are mounted on the outside of the mounted on the front panel. On the rear is the coax output connector. Four rubber bumpers are mounted on the bottom to prevent scratches from the hardware which fastens the parts to the enclosure.

When assembling the parts on the main PC board, a socket or Molex pins should be installed at U11 for the integrated circuit. It may be necessary to select a 7404, since its upper frequency limit is being pushed. But at the current price of 25 to 35 cents each, it is much more economical to buy three or four than to buy the devices which would be required for a more sophisticated Schmitt trigger. More about this later.

The main board is connected to the

frequency-control switches from 20 pads, designated A through V, along the front of the board, as shown in fig. 5. The remainder of the connections appear in fig. 8. All connecting leads to and from the board should be soldered in place on the board before connecting the other ends. Leave plenty of wire on each lead to make the connections to the parts which are not on the board so that if any troubleshooting must be done it will not be necessary to unsolder the wires. Shielded wiring is made with RG-174/U coax. All wiring, except for that between the "VCO" pad and E1 on the vco, carries dc only, making lead length noncritical. The board is mounted on four standoff posts; in order to prevent ground loops, three of the four should be non-metallic or should be insulated from the ground plane on the board, so that only a single metallic post grounds the main board to the cabinet.

The vco printed-circuit board is enclosed in a 2-1/8x2-5/8x2-3/4 inch (5.4x6.7x7 cm) mini-box. Feedthrough terminal E1 and feedthrough capacitors C17, C23, C24 and C33 are mounted on one side of the box, and J1 is placed on one end, corresponding to the leads designated in fig. 12. Use solder lugs under the outsides of the feedthrough capacitors which are to have shielded leads attached. The PC board is fastened to the side of the box on which the feedthroughs are mounted by means of small right-angle brackets which are soldered to the ground plane of the board. The feedthroughs and J1 are then connected to the appropriate pads on the board, using short lengths of bare wire.

The vco mini-box must be mounted so that there is only a single return to common ground, in order to prevent ground loops. When mounting the vco,

table 1. Semiconductor and miscellaneous parts list.

CR1-CR24	1N34A, 1N100, 1N270 or equivalent germanium diode
CR25	Motorola MV1401 varactor
CR26,CR27	1N658
CR28	1N914 or 1N4148
CR29-CR32	1N4001 or equivalent 50 PIV, 1 amp silicon rectifier
CR33	Hewlett-Packard 5082-4882, Motorola MLED655, Radio Shack 276-041 or equivalent light-emitting diode
E1	insulated feedthrough terminal
L1	5-3/4 turns no. 28, closewound on 0.211" (5.5mm) diameter slug-tuned form (Miller 25A014-4)
L2	10-3/4 turns no. 28, closewound on 0.211" (5.5mm) diameter slug-tuned form (Miller 25A014-3)
L3	20 turns no. 30, closewound on 0.211" (5.5mm) diameter slug-tuned form (Miller 25A014-3)
Q1	Siliconix E300, 2N5397 or 2N5398
Q2	2N2219
Q3	2N5458, 2N5459 or Motorola MPF103
S1	2-pole, 5-position, non-shorting rotary switch
S2	3-pole, 10-position, non-shorting rotary switch
S3	4-pole, 3-position, non-shorting rotary switch
U1	Motorola MC846P quad 2-input NAND gate
U2	7490 decade counter
U3,U4	7474 dual D-type flip-flop
U5	National LM3900 quad op amp
U6	74S196 or Signetics 82S90 presettable decade counter
U7,U8,U9	74196 or Signetics 8290 presettable decade counter
U10	7430 8-input NAND gate
U11	7404 hex inverter (see text)
U12	74S112 dual J-K flip-flop
U13	7812 or National LM340-12 12-volt regulator
U14	7805, National LM340-5 or LM309K 5-volt regulator

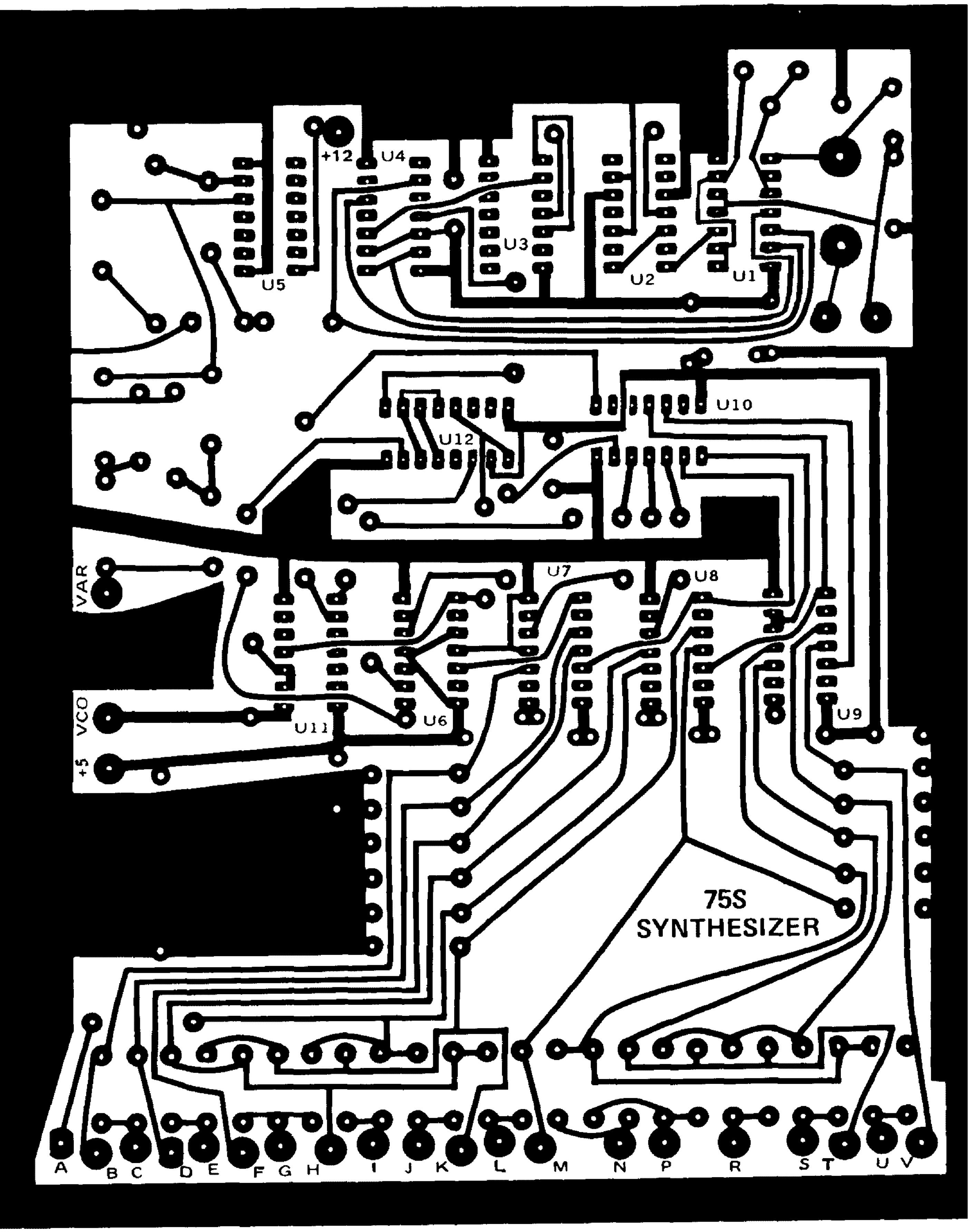


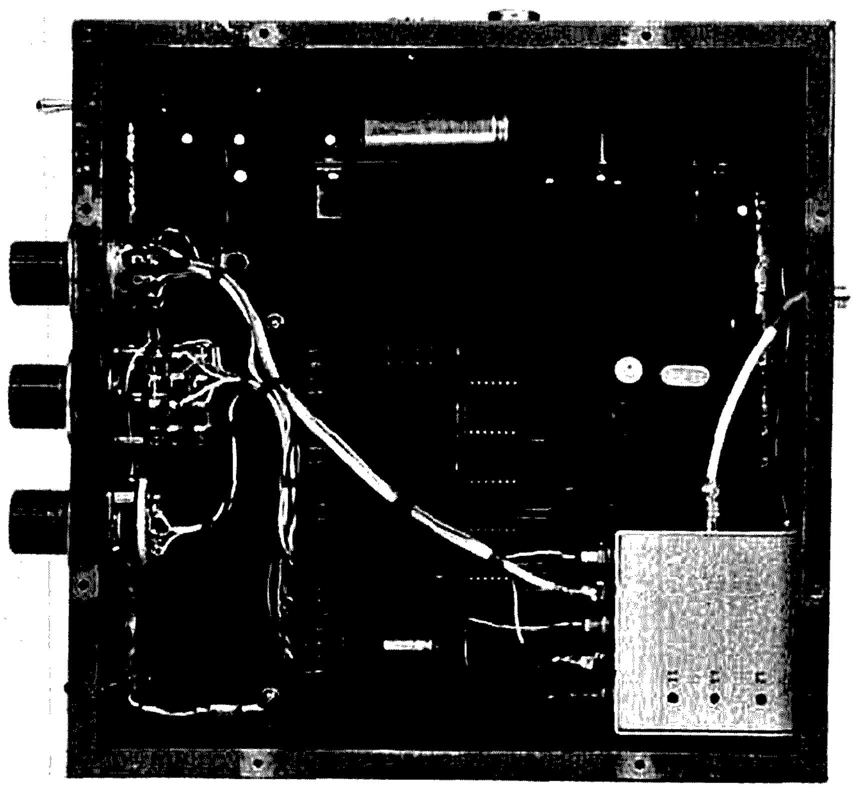
fig. 9. Foil pattern of the main printed-circuit board.

be sure to position it so that the feedthroughs are close to the side of the main PC board on which are located the pads designated "VAR" and "VCO." The lead between E1 on the vco and the "VCO" pad on the main board should not be more than 3 or 4 inches (7.5 or 10 cm) long. The shields of the leads going to capacitors C17, C23 and C24 are grounded to the solder lugs under those capacitors.

The coax between J1 and the output connector on the rear of the cabinet should be 95-ohm type, such as RG-180/U or RG-195/U (although using a very short piece of 50-ohm RG-174/U will probably have little detrimental affect). The cable between the synthesizer and the receiver *must* be the 95-ohm type or 93-ohm RG-62/U.

The power transformer should supply 16 to 17 volts ac at approximately 0.5 ampere. Any higher voltage only results in greater heat dissipation in regulator U13. The method by which this voltage can be obtained by modifying an inexpensive 24-volt transformer is described in reference 1.

The numbered positions on the rotary switch knobs were made by using number transfers on the skirts of the knobs. Several heavy coats of Krylon



Interior of the assembled synthesizer. The power supply components are located to the left of the main PC board (top). The vco is mounted on the rear wall just above the main board (bottom right). Note the holes in the vco enclosure which provide access to the tuning slugs of the coils mounted on the vco board inside.

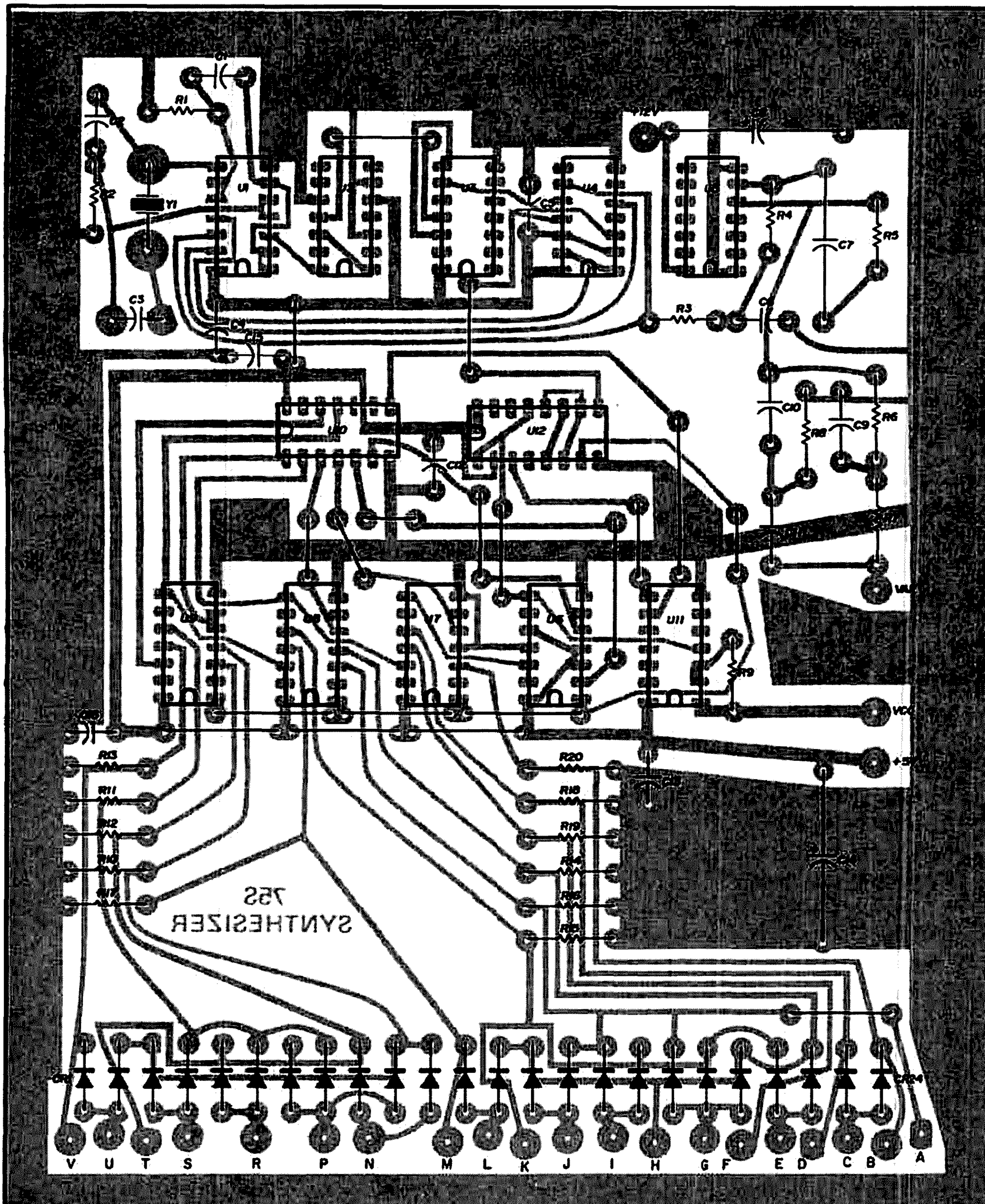


fig. 10. Location of parts on the main printed-circuit board.

fixative were sprayed on to keep the numbers from peeling off. The panel markings can be applied in the same manner. If you are fortunate enough to have access to a Selectric Composer, or even a good electric typewriter, the panel labeling can be typed on frisket, which is an adhesive-backed translucent

acetate. The material is virtually invisible when applied to a grey panel, against which the black type effectively contrasts. Frisket is available at art-supply dealers.

receiver modifications

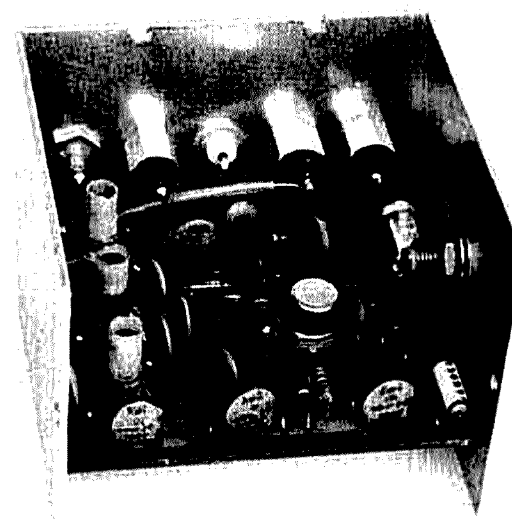
At the beginning of the article, I indicated that only minor modifications had to be made to the receiver; these are shown in **fig. 13**. The synthesized crystal frequency is introduced into the receiver via the *spare* jack on the rear apron. The added components must be placed close to the oscillator-mixer (V3) tube socket, and the shielded connection to the *spare* jack made with 93- or 95- ohm coax, e.g. RG-62/U, RG-180/U or RG-195/U.

The existing coax lead in the receiver between *bandswitch* S1 and pin 2 of V3 is disconnected from V3 and rewired to the added relay. This relay may be a reed or "crystal can" type; its coil resistance must be at least 500 ohms in order to minimize the voltage drop across R32 in the vco.

When the synthesizer is off or is disconnected from the receiver, the normally-closed relay contacts complete the circuit between S1 and V3, allowing the receiver to function as if no changes had been made. When the synthesizer is turned on, the relay is energized by the 12-volt supply in the vco through R32 and L8 (see **fig. 4**), disconnecting the crystals and applying the synthesized crystal signal to the control grid (pin 2) of the oscillator section of V3. The 56-μH choke at the relay coil isolates it from the rf signal path.

A new *preselector* scale may be added to the receiver so that you don't have to consult or memorize the preselector chart in the manual. **Fig. 14** is a full-scale reproduction of the new *preselector* scale and may be cut out, or photocopied if you prefer not to mutilate the magazine.

Remove the *preselector* knob and pointer and attach the scale to the front panel of the receiver, using either a spot of rubber cement or a small piece of double-sided sticky-back tape in each corner. The scale shows the approximate settings of the *preselector* control; the letter at the end of each scale segment indicates the *bandswitch* position to be used.



Construction of the vco sub-chassis. Inductors L1, L2 and L3 are at left. Output connector is at right. Feedthrough capacitors and input connector are on rear panel. All other components are mounted on the printed-circuit board at bottom.

alignment and test

After all wiring and connections have been checked and rechecked, the synthesizer is ready for the few adjustments necessary to set the vco on frequency. The only test equipment absolutely necessary is an electronic voltmeter, although a frequency counter and oscilloscope can be helpful.

Apply power to the synthesizer and

check the supply voltages to make sure that they are within five percent of the nominal values. Then make sure the reference oscillator is working by bringing a lead from the receiver antenna jack close to the crystal. Harmonics of the 100-kHz oscillator should be heard in the receiver. The oscillator may also be checked with a scope; 100-kHz square waves should be observed at pin 3 of U1, and 5-kHz square waves should be present at pin 9 of U3. If the oscillator is not working, adjust trimmer capacitor C3, although it is not necessary for the crystal to be oscillating at exactly 100 kHz at this time.

Connect an electronic voltmeter between ground and feedthrough capacitor C17 on the vco. Set the rotary switches to 9.8 MHz and adjust the tuning slug of L3 until a meter reading of 10 volts is obtained. Rotate the *units* switch toward zero, noting that the voltmeter reading drops with each change in the switch setting until position 3 is reached. Then rotate the *tenths* switch to zero, noting that the voltage continues to drop. (Actually, any switch setting below 3.4 MHz is invalid, since it is below the tuning range of the receiver preselector.) A frequency counter connected to the output of the synthesizer should indicate approximately 12.955 MHz with the frequency switches set to 9.8 MHz, and 6.155 MHz with the switches set to 3.0 MHz. The exact frequencies will be obtained only if the crystal is set to exactly 100 kHz, which adjustment is not made until the vco is aligned.

Next, set the rotary switches for a frequency of 19.8 MHz, and adjust the slug in L2 for a voltmeter reading of approximately 10 volts. *Do not touch the slug in L3*. Again turn the *units* switch toward zero and note that the voltage drops at each switch position, including position 0. Turn the *tenths* switch to 0 and make sure that the voltage also decreases with each step. A counter should indicate about 13.155 MHz and 22.955 MHz respectively for the minimum and maximum *units* and *tenths* switch settings.

Finally, set the rotary switches for a frequency of 29.8 MHz and adjust L1 as described in the preceding adjustment. The frequency range of the vco with the *tens* switch in position 2 is 23.155 to 32.955 MHz. If you find that it is not

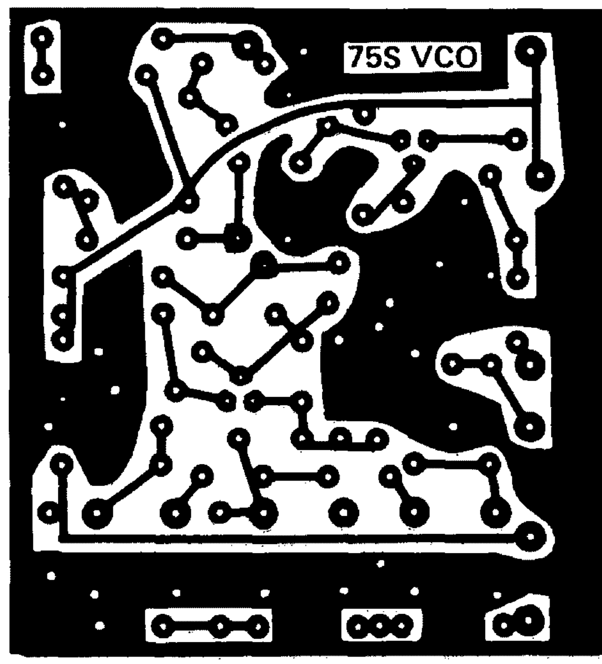


fig. 11. Foil pattern of the vco printed-circuit board.

possible to obtain a 10-volt reading on the voltmeter at switch settings of 29.8 MHz, but that at some lower frequency the loop starts to lock up (as indicated by incremental changes in the voltage as the *tenths* or *units* switch positions are changed), you must change U11. I tried four 7404 hex inverters in the circuit; two worked at the highest frequency and two quit at about the 26-MHz switch settings. Thus there is every probability of getting at least one better-than-average IC out of three or four.

Connect the synthesizer output to the *spare* jack on the receiver, using a cable made from 93- or 95-ohm coax,

Turn on the receiver and synthesizer and set the synthesizer switches to one of the WWV frequencies suitable for good reception. Set the receiver *bandswitch* and *preselector* control to the settings specified for the frequency selected. WWV should be heard when the receiver is tuned to zero. Then allow the synthesizer and receiver to warm up for about a half-hour.

Wrap a few turns of one end of an insulated wire around V2 in the receiver and bring the other end close to the 100-kHz crystal in the synthesizer. Carefully adjust the crystal trimmer capacitor in the synthesizer to zero-beat with WWV. It may be necessary to retune the receiver slightly when doing this because changing the frequency of the crystal changes the synthesizer output frequency, which is now the receiver hfo frequency.

That completes the alignment – you now have a general-coverage receiver.

Operating the receiver with the synthesizer involves no more than turning on the synthesizer and setting its switches to the desired frequency. Then set the receiver *bandswitch* and *preselector* control to the appropriate positions and tune.

One minor difference will be noted when using the synthesizer. On the 10-meter band, the 200-kHz segments start at 28.0 MHz and progress in 200-kHz increments so that you tune from 28.4 MHz, 28.6 MHz, etc. In other words, while each segment throughout the synthesizer range will start with an even *tenths* digit, the crystals supplied with the receiver set the 10-meter segments at 28.5, 28.7, and 28.9 MHz.

There is one precaution which *may* be necessary. Even with the prototype unit enclosed in a steel box, there was still a very small amount of 60-Hz pickup from the receiver transformer when the synthesizer was placed next to either side of the receiver. (The condition also exists with the synthesizer on top of the receiver, but this arrangement would block the heat convection flow

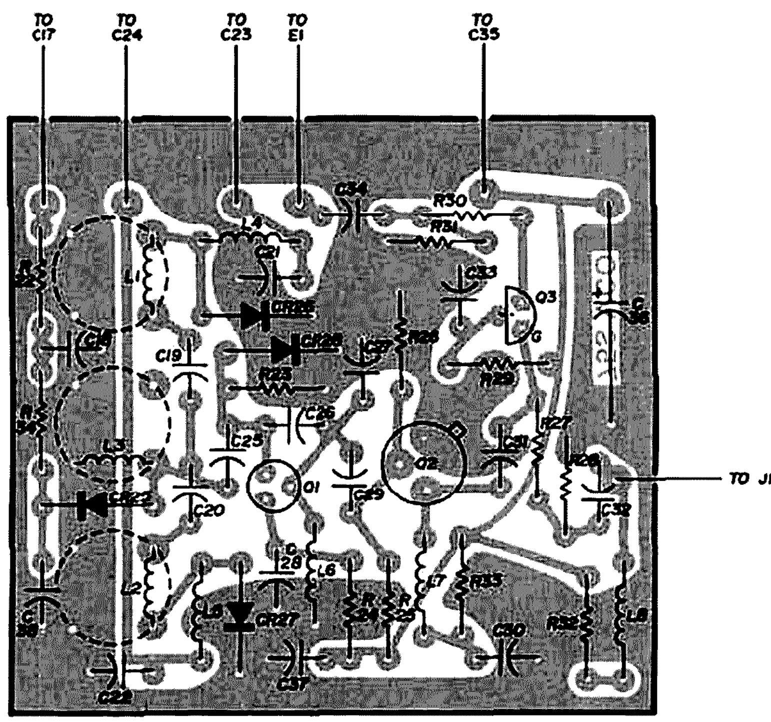


fig. 12. Location of parts on the vco printed-circuit board.

from the receiver and should not be used anyway.) Simply moving the synthesizer 3 to 4 inches (7.5 to 10 cm) away from either side eliminated the stray pick-up. It is entirely possible that this condition may not manifest itself with all receivers, and it will undoubtedly *depend* on *the physical locations of* the assemblies within the synthesizer cabinet. In any event, the synthesizer is still close enough to the receiver for convenient operation.

conclusions

The synthesizer satisfies all the requirements necessary to make any Collins 75S a general-coverage receiver. Spurious signals are down a minimum of 80 dB on all frequencies, and are down better than 90 dB on most. The major spurs appear 10 kHz either side of the incoming signal, and are caused by the second harmonic of the 5-kHz reference frequency. The reference frequency itself is weaker than the harmonic because of the attenuation provided by the parallel-T filter. Although the suppression of the spurious sidebands was achieved at the expense of fast lock-up time, a one- or two-second lock-up is of little consequence, since it takes that long to move your hand from the synthesizer switches and tune the receiver.

The synthesizer has not been used with a 75S receiver operating in transceive mode with a 32S transmitter, although there is little reason to doubt that it will work. There should be sufficient sideband attenuation to keep spurious outputs from the transmitter at least 60 dB down. The only possibility of trouble might be rf getting back into the synthesizer from the transmitter, which would be simply a shielding problem. However, until and if new amateur bands are forthcoming, there is no reason to use the synthesizer for transmitting except possibly on part of the 10-meter band.

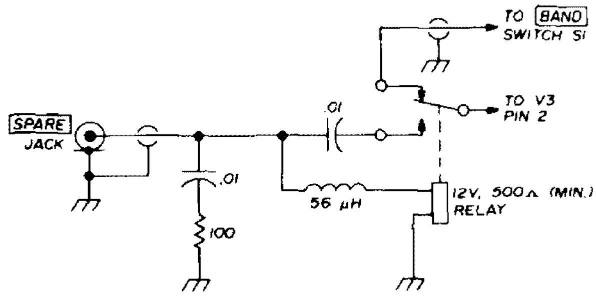


fig. 13. Wiring changes to be made to the Collins 75S receiver. Make sure that the relay will pull in when it is connected in series with a 110-ohm resistor to a 12-volt dc supply.

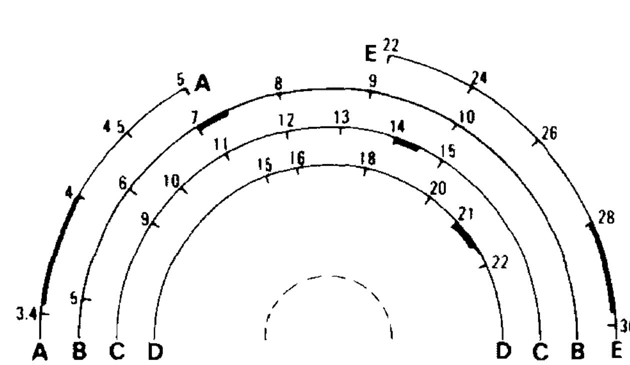


fig. 14. New receiver preselector scale for the Collins 75S receiver.

acknowledgements

The following were instrumental in enabling me to complete this project; without their assistance this article would never have been written: Cliff Buttschardt, W6HDO, for the use of his 75S-3 receiver; Duke Moran, W6SPB, who etched the prototype PC boards; Paul Zander, WB6GNM, for his invaluable help in the design of the phase-locked loop; and Bob Melvin, W6VSV, who listened to my problems and even made a suggestion or two.

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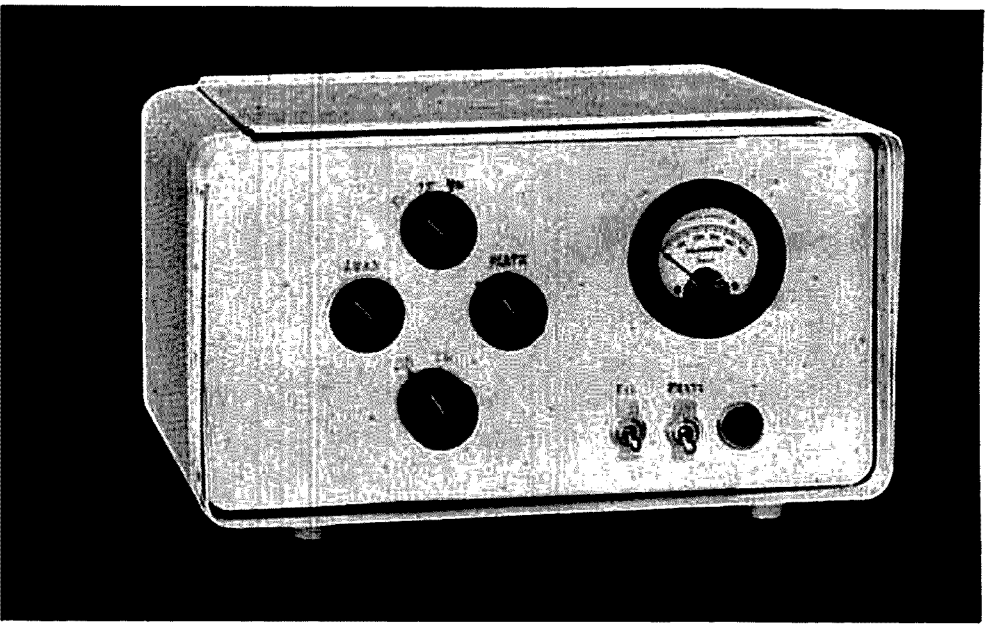
ham radio



100-watt linear amplifier for QRP rigs

A compact rf amplifier for ultra low power amateur transmitters

W. S. Skeen, W6WR*

To dispose of the first question apt to be asked – "Why add a linear amplifier to a QRP rig since it defeats the whole idea of QRP operation?" – I'd like to say that while 2 to 5 watts can do wonders during the day, night operation is a different story. The prevailing sunspot activity precludes the predictable propagation conditions of 10 or 15 years ago, even on the 40- and 80-meter bands. A little boost to the output of a QRP rig means the difference between fun and drudgery during nighttime operation. When good conditions return, a linear amplifier probably won't be needed for low-power work.

I built this amplifier to augment a homebrew rig that didn't live up to my expectations – the rig had only about 5 watts output on ssb. The linear amplifier described here should prove to be a useful adjunct to low-power transmitters in the 2 to 10 watt range.

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circuit

The schematic (fig. 1) is simple, effective, and uncluttered. A grounded-grid, grounded screen circuit is used without the usual blower for tube cooling. Any of the Eimac 4X or 4CX tubes will perform equally well.

In the late 1950s, I witnessed some tests at Eimac that were run to see what these tubes would do without forced air cooling. The tests indicated that the tubes would dissipate 60 to 70 watts under key down operation – however, the tubes were mounted *in the open*, with no restriction to ambient air flow. Under intermittent operation, it appears that a tube of this family could safely dissipate at least 50 watts; perhaps a little more if cooling fins are provided. Again, the qualifier is: unrestricted air flow around the tube. A further advantage is that the heater-cathode isolation in these tubes is excellent; no filament chokes are required at this power level.

One other precaution should be observed. Although the heater is rated at 6.0 volts ±5%, it is recommended that 6.0 volts be considered the upper limit. A 50-ohm, 10-watt resistor in series with the primary of a 6.3 Vac, 2.5 amp filament transformer should do the trick. The heater contributes a large part of the heat to be dissipated, and tube life is prolonged by keeping the heater voltage on the low side.

A noninductive, 1-watt carbon resistor (R1) of a few hundred ohms is provided for situations where excitation is excessive with no provision for reducing it. The resistor should be selected to obtain recommended operating conditions. While this amplifier is a two-band affair for 80 and 40 meters, additional taps can, of course, be provided for other bands on the pi-net output coil.

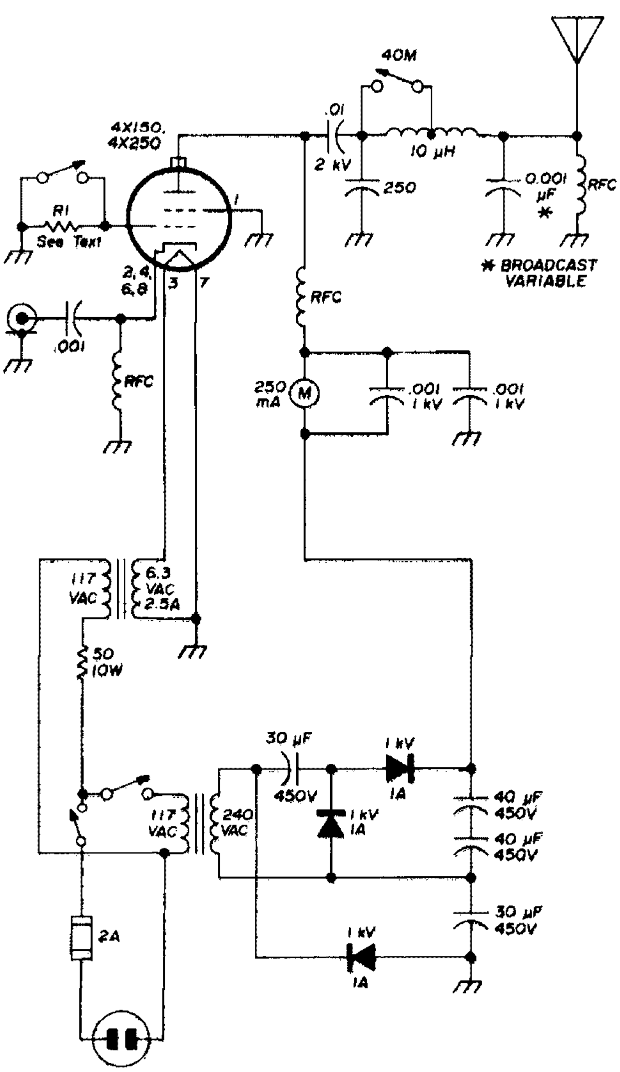


fig. 1. QRP linear schematic. Any of the 4X or 4CX series of tubes may be used. Up to 50 watts dissipation is possible without a blower providing tube is mounted in the clear.

operating conditions

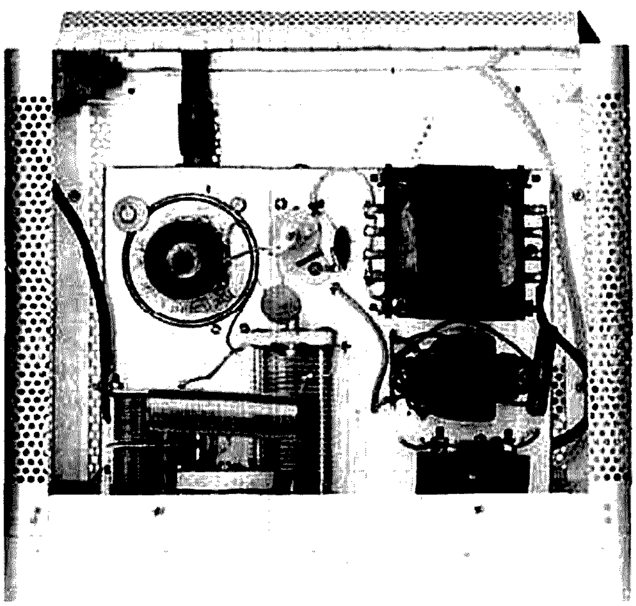
The amplifier has a power gain of about ten with both grids grounded, so 5 watts input should yield about 50 watts output, with a plate current of 100 mA and E_b at 1 kV. This current is 100 mA as read on the meter in the CW mode. Although up to 150 mA can be obtained with 7 watts input on CW, it is recommended that the series grid resistor be switched in to hold the plate current to 100 mA on ssb or 150 mA on CW. As with any low duty cycle amplifier, don't hold the key down longer than necessary.

Static plate current (no drive) is about 10 mA. Linearity could be improved by a higher idling current, but observations with a spectrum analyzer indicate that, with 10 mA static plate current, the bandwidth is entirely

acceptable, and reports have been universally good.

construction

An aluminum chassis, 2x6x9 inches (5x15x23cm) in an LMB cabinet constructed with perforated aluminum for sides top and bottom, easily contains the amplifier with power supply. The



Chassis layout. Power transformer (upper right) is a 50VA isolation transformer. Note air space around tube.

LMB cabinet (model CO-2) measures 6½ inches high by 10 inches deep by 13 inches wide (16.5x25x33cm) excluding hood. The chassis was purposely selected to improve air circulation. The tube socket is mounted as close as possible to one rear corner, both for short leads from coax connectors *and again* to improve cooling by air circulation over the chassis edge (see photo).

While the built-in screen bypass capacitor of the SK-600 socket is superfluous, open space around the tube pins permits air flow around the header (tube base). Old timers will remember that these tubes also fit in a loctal socket. A 4-inch (10.2cm) square piece of perforated aluminum with chassis cutout should also be suitable.

The grid resistor shorting switch, if used, should be mounted close to the tube, which means a shaft extension. The rf chokes are *ordinary* garden-variety 2 mH, 100 mA chokes with the exception of the plate choke; but even here, tests indicated that the smaller chokes should hold up.

The 10-μH pi-net coil (46 turns, 7/8 inch [2.2cm] OD, 3 inches [7.6cm] long, air wound) is barely large enough to cover the low end of 80 meters and is tapped at slightly less than one-half for 40 meters. The coil is mounted on the switch. A paralleled BC tuning capacitor suffices for the output. Additional fixed capacitance to total 1000 pF can be used, if necessary, for 80 meters.

power supply

A 50-VA isolation transformer is used with 120 Vac input and 240 Vac output to a voltage tripler arrangement that provides the 1-kV plate supply. Inspection of recent catalogs indicates that 115/230 V primary, 115 V secondary transformers are about all that are available now (about $10). In this case, the secondary may be used for the primary with a slight loss in output voltage and regulation. The filter capacitors and diodes mount under the chassis, and since there isn't much else there, no under chassis photograph is provided. A separate heater transformer and switch are provided. The tube heater should be allowed to warm up at least a half minute before applying plate voltage. Static plate current provides "bleeder" protection.

summary

This small linear compares favorably, both in size and performance, with commercially built units of the same power class. It has held its own with other 200-watt-plus units and has provided many solid contacts during the worst interference hours of the evening.

ham radio

an introduction to microprocessors

David Larsen, WB4HYJ, Peter Rony and Jonathan Titus*

Microprocessors are probably the single, most exciting development in the entire field of electronics, and in this article, the first of a series on microprocessors, we would like briefly to compare them to programmable calculators for typical laboratory applications.

The best description of what a microprocessor is, and isn't, was given by Laurence Altman in a recent issue of *Electronics*:[1] "A microprocessor is not a computer but only part of one. To make a computer out of a microprocessor requires the addition of memory for its control program, plus input and output circuits to operate peripheral equipment . . . What a microprocessor is, then, is the control and processing portion of a small computer or microcomputer. Moreover, it has come to mean the kind of processor that can be built with LSI mos or, more recently, bipolar circuitry, usually on one chip. Like all computer processors, microprocessors can handle both arithmetic and logic data in bit-parallel fashion under control of a program. But they are distinguished both from a minicomputer processor by their use of LSI with its lower power and costs, and from other LSI devices (except calculator chips) by their programmable behavior."

Thus, a microprocessor is not a totally self-contained computer-on-a-chip, nor is it able to complete with and replace the central processing unit (CPU) within a computer. Existing microprocessor chips are simply much too slow for such applications. The niche that microprocessors will soon fill is in the creation of "smart" input/output devices to a computer that relieve the computer of the drudgery associated with the data acquisition from and the control of such devices. In other words, microprocessors will shortly become very important tools in computer interfacing, a trend that will accelerate as the price of microprocessor chips declines, as more individuals develop the capability to handle such chips, and as more manufacturers incorporate such chips in laboratory instruments and other types of devices that communicate with computers.

The advantages of interfacing with microprocessors are at least fourfold:

1. Microprocessor communications are simple. The communications capability of a microprocessor system is a big point in its favor. Most such systems come with a built-in asynchronous serial port, and thus can communicate with teleprinters or with any device that also has an asynchronous serial port. The microprocessor is not inherently limited to only a single asynchronous port; it is

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very easy to add more such ports and thus permit the microprocessor system to communicate serially with other external devices such as laboratory instruments that are interfaced with Analog Devices' *Serdex* modules.

Microprocessor systems have parallel input ports for inputs from various digital sensor instruments, including voltmeters, panel meters, frequency meters, and counters. Any type of digital circuit that can supply parallel digital data can be used in conjunction with a microprocessor system.

2. Microprocessor systems are inexpensive. Such systems currently range in price from several hundred dollars to several thousand dollars, depending upon the capability of the system. They are available from Intel, Prolog Corporation, E & L Instruments, Control Logic, and other companies. The number of manufacturers that offer microprocessor systems is increasing rapidly.

3. Microprocessor systems are flexible and powerful. Microprocessors have the ability to make decisions. (Is an input value from a digital sensor too high or too low? If it is too high, then open a valve and release pressure on the system. If it is too low, then open another valve and add gas to the system.)

Microprocessors use software to replace hardware; i.e., microprocessor programs replace complicated hard-wired random logic digital electronic circuits that perform a variety of functions, including sequential logic, non-sequential logic, simple arithmetic calculations, and comparison of digital signals. Manufacturers of microprocessor systems provide you with both read/write memory, for temporary data and program storage, and with read-only memory, which is easily programmed with the aid of a PROM programmer. Once you have written and tested a program using read/write memory that can acquire data and perform desired control operations, you can "burn" it into a programmable read-only memory (PROM) IC and then use that chip day after day to operate the microprocessor system.

You never have to worry about a power failure causing your program to be erased. The program can remain in the PROM for up to twenty years; it is always available for reloading into a read/write memory. The program can be easily modified to accommodate changed data acquisition or control requirements. You can develop a whole repertoire of PROM chips to accomplish different functions.

4. Microprocessor systems are capable of handling most laboratory data acquisition requirements. Current microprocessor systems can acquire digital data at the rate of five hundred 16-bit words per second. Higher data acquisition rates are occasionally claimed by manufacturers, but they frequently overlook the real software overhead that is needed, for example, to input the data, check if the data are ready, and compare the data to make sure that they are within the right range of values.

In the area of mathematical computations, microprocessors can perform integer multiplications and divisions, i.e., 3 times 4 or 5 divided by 7, with reasonable accuracy. A floating-point package available with the 8-bit Intel microprocessor allows you to perform additions, subtractions, multiplications, and divisions over the range of $\pm 10^{32}$ to $\pm 10^{-32}$. This package requires four read-only memories, which means that 1000 words of your microprocessor are

This the first of a new series of articles on the subject of microprocessors which we will be presenting in future months. Material presented here is reprinted with permission from *American Laboratory*, June, 1975, copyright © International Scientific Communications, Inc., Fairfield, Connecticut, 1975.

dedicated to the floating-point package. Execution times are slow, so you must worry about the following types of questions: Do you acquire a data point and then operate upon it and still have sufficient time to acquire the next data point? Or must you store a complete block of data and then operate upon the block as a whole? If you store a block of data, how much additional memory is required for the microprocessor? Finally, is the system sufficiently complex and expensive that it can be replaced by a minicomputer or programmable calculator?

The strong point of the microprocessor is that it can perform control functions quickly, easily, and inexpensively. The microprocessor can turn devices on and off. It can regulate physical parameters such as temperature, pressure, velocity, and flow. Since it lacks special functions such as log, χ^y, sine, cosine, square root, hyperbolic sine, and hyperbolic cosine, it cannot perform sophisticated mathematical computations. This is one reason why many individuals are looking very seriously at programmable calculators, which start in the vicinity of $3000; are available from Wang, Tektronix, and Hewlett-Packard; and allow the user to program with complex functions such as sine, cosine, log, and χ^y. The programmable calculators, however, are not nearly as convenient to use as microprocessors in the control of equipment and processes.

As a final point, we would like to caution you about making any long-term decisions concerning both microprocessors and programmable calculators. The comments above apply to today's technology, which is precisely what you can do today. The price/performance ratio changes from day to day so that a decision that is valid today may not be the same one that would be proper in a month or a year from now; e.g., 8-bit bipolar microprocessors now available from Intel have cycle times of 50 nanoseconds. This speed is a little bit difficult to precisely define for the user, but it represents probably a decade of improvement in overall microprocessor speed when compared to any microprocessor available a year ago.

If you can postpone your problem, you may find that you can solve it differently and/or less expensively a year from now. Digital electronics is without doubt the fastest changing technological field today. You, as an amateur, engineer or scientist, will be a major beneficiary of the changes that are occurring. However, to take proper advantage of the new technology, you will have to spend some time learning the jargon and understanding the tradeoffs that can be made.

Microprocessor equipment, if cared for properly, has an operational life of at least ten years but a functional life that may only be several years. A reasonable strategy would be to postpone the purchase of a microprocessor until the price/performance ratio justifies a purchase, and then to go ahead and purchase a system with the knowledge that the same system will probably cost at least 20% less for the same performance a year later. We believe that not too much time will pass before all of us who are involved in research or manufacturing and depend upon instrumentation will have to take advantage of the power of microprocessors if we are to continue to have viable products or research programs.

We recommend that you give careful consideration to the ability to interface newly acquired digital instruments to future ones that will come on the market within the next several years. We emphasize again that the existence of asynchronous serial ports on your digital instruments will allow you to hedge your bets for the future.

reference

1. L. Altman, "Single-chip microprocessors open up a new world of applications," *Electronics*, April 18, 1974, page 81.

ham radio

squelch circuits

for transistor radios

Agc-activated squelch can easily be added to portable transistor radios

Robert Harris, Jr., WB4WSU, 801 Tucson Road, Virginia Beach, Virginia 23462

Inexpensive transistorized portable radios can become excellent monitor receivers for vhf operators (a-m and fm) with the addition of one of the simple squelch circuits presented in this article. The squelch will get rid of the constant and fatiguing hash and noise usually put out under no-signal conditions, making the portable a much more useful and enjoyable radio to listen to. I have had much success in adding the circuits shown here to several portables. It should be possible to adapt this same basic approach to just about any existing portable receiver.

A brief explanation of squelch circuitry seems appropriate at this point. During no-signal conditions, the audio output of a receiver is random and unpleasant noise. When a signal is received, the gain of the receiver is reduced or limited by agc action, and the level of noise output is reduced. The greater the signal strength, the lower the noise output, and hence the term "receiver quieting," By adding circuitry which detects the degree to which the receiver has been quieted, and by using this circuitry to mute or un-mute the audio output, squelch may be added to the receiver.

Note, that with the squelch I have just described the operator may adjust the sensitivity of his receiver in terms of a minimum signal-to-noise ratio needed to produce an audio output signal. By making this signal-to-noise ratio sufficiently high, the operator can be sure that whenever the receiver produces an audio output it will contain a signal of a

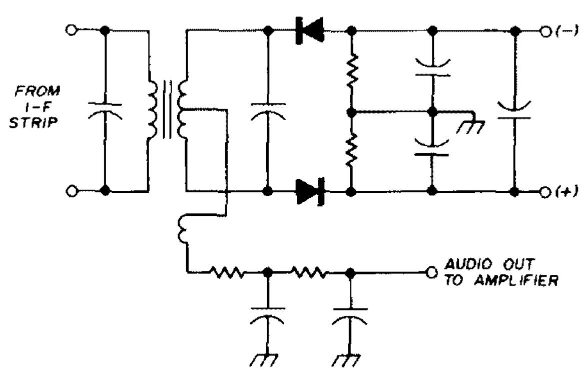


fig. 1. A typical ratio detector found in many transistor radios. Agc voltage may be taker from the (+) or (−) terminals.

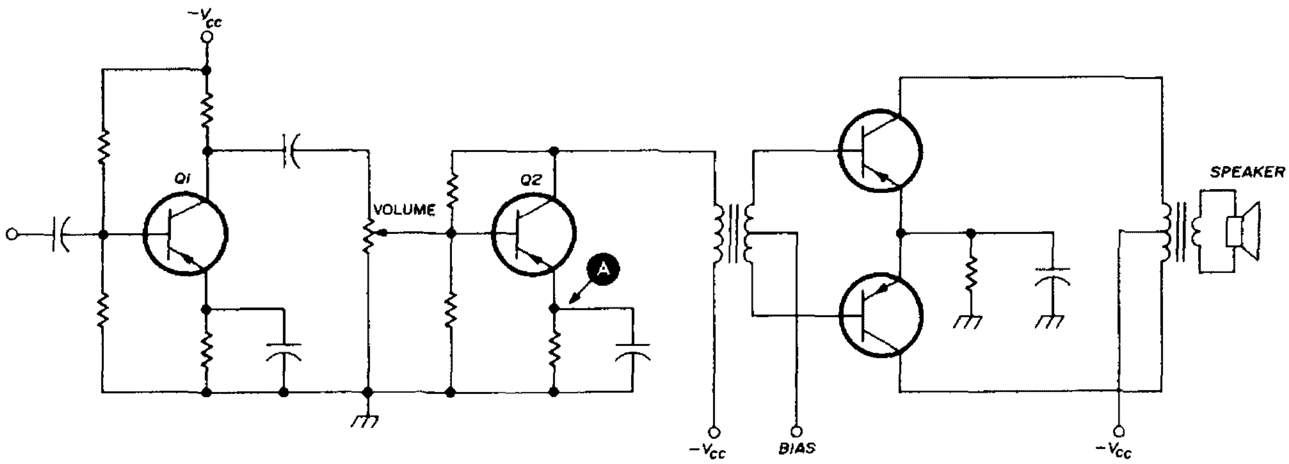


fig. 2. The audio section of most portables will closely resemble the circuit shown here. Point A in Q2's emitter is a control point for muting the audio output.

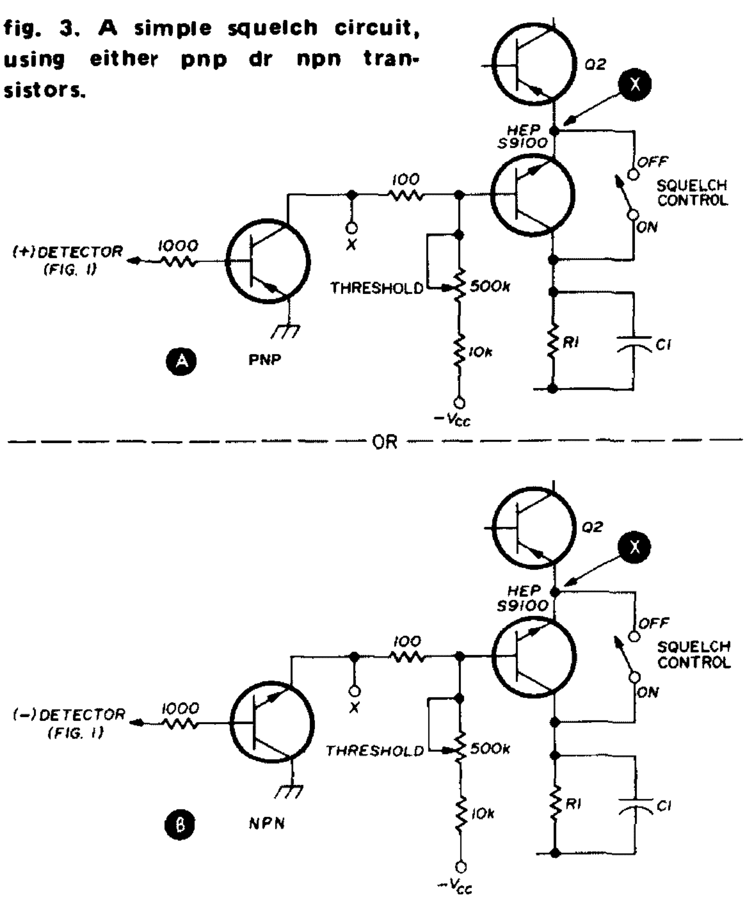
certain minimum readability. At first this might sound like intentionally reducing the sensitivity of your receiver, but this is not so. With sophisticated circuits, the opening of the receiver squelch alerts the operator to the presence of marginal level signals that might otherwise have gone unnoticed in the noise. Unfortunately, these noise-operated squelch circuits are somewhat complex, and they are beyond the scope of this article.

Another method of producing squelch action is to make use of the agc or other signal level dependent voltages to control audio muting. The only drawback with this approach is that sufficiently strong noise or interference will also open the squelch. However, for the purpose intended here they work quite acceptably.

detector

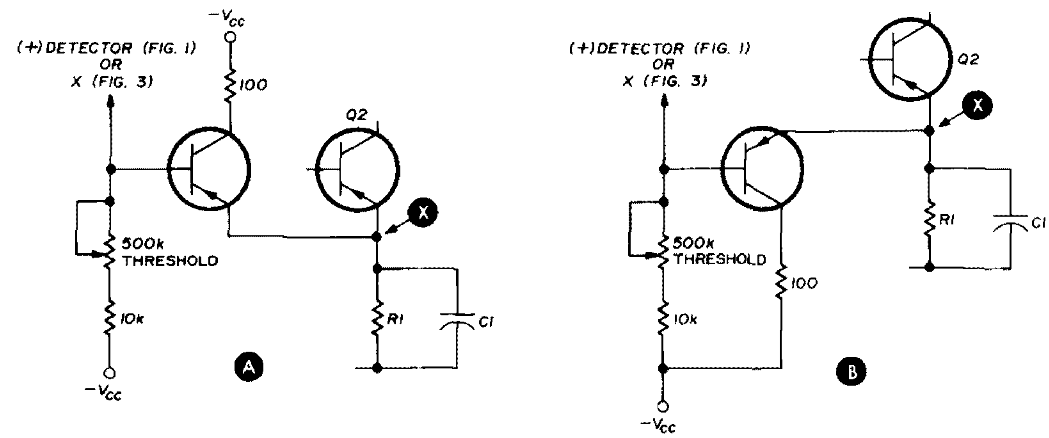
Most inexpensive portable receivers use the common ratio detector similar to the one shown in fig. 1. This detector develops significant positive and negative voltages during signal conditions, either of which may be used for control of the audio muting. Generally, the audio section of a portable receiver will closely resemble the circuit of fig. 2. Audio from the detector is coupled to Q1, a preamplifier, (which is sometimes omitted in inexpensive sets). The output is coupled through the volume control

fig. 3. A simple squelch circuit, using either pnp dr npn transistors.



to the driver stage, Q2, which in turn drives the output stage. These sections are easily located by finding the audio transformers associated with them, and it is seldom necessary to resort to a schematic to find the desired stages.

plete muting are shown in **fig. 4**. In these circuits, the emitter signal path of the driving transistor is not broken, but the biasing of the driver is upset when no signal is present. But with even a small signal present, the bias is sharply



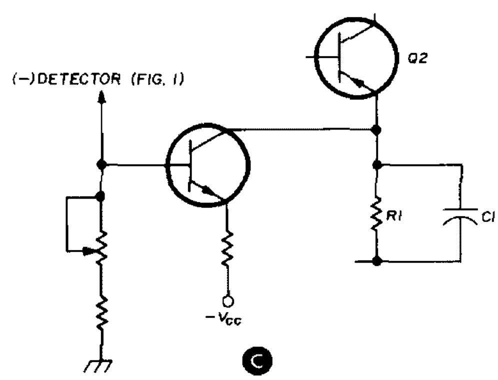


fig. 4. A "brute force" squelch circuit is shown at (A); variations of the circuit are shown in (B) and (C). A spst switch may be placed in the control line for on/off control of the squelch.

simple squelch

The point marked X in the emitter of Q2 is a convenient point to add squelch control to the audio stages. By breaking the circuit at this point and adding the muting circuit of **fig. 3A**, a simple squelch circuit is obtained. In some cases, small amounts of signal or noise will leak through even when this stage is supposedly squelched. Also, an increase in distortion may be noticed at high volume levels. A variation of this circuit is also shown, using an npn transistor, in **fig.** 3B.

improved design

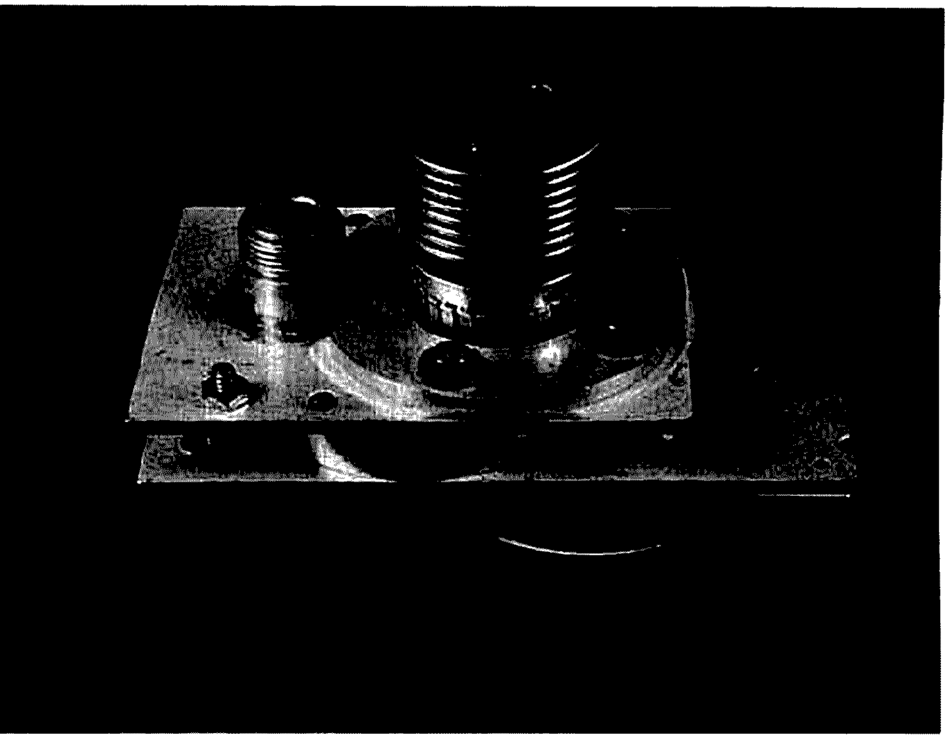
Somewhat more positive acting squelch circuits which do not tend to cause distortion or suffer from incom-

returned to normal and audio output is restored.

conclusion

The several approaches to agc-controlled squelch shown here can be easily adapted to most portable radios. Receivers using both npn and pnp transistors may be accommodated, and junk box transistors seem to work a great percentage of the time. If you are unable to locate a source of agc voltage, do a little poking around with a vtvm until you find a voltage source that varies with signal strength. That is all that is needed to add squelch to a portable radio, making it a much more useful and enjoyable low cost monitor receiver.

ham radio



1152- to 2304-MHz power doubler

Construction of a single-tube frequency doubler for 2304 MHz that provides 5 dB power gain

Norman J. Foot, WA9HUV*

A 2C39 power **amplifier** capable of providing 30 watts output on 2304 MHz was described in the February, 1975, issue of *ham radio*.[1] Its design involves a combination of the best characteristics of various experimental 2.3 GHz amplifier models I have built and tested over the past three years. Each succeeding version differed from the previous one in ways which both improved performance and simplified construction.

During the same three-year period, on-the-air tests over a ten-mile path were performed between WA9HUV and both W9DCN and K9CNN, first working 432/2304 crossband, and later using 2304 MHz two-way. Signals were well over S9 on 2304 MHz, in spite of the 1296-MHz antenna used by W9DCN. Contacts over longer distances have

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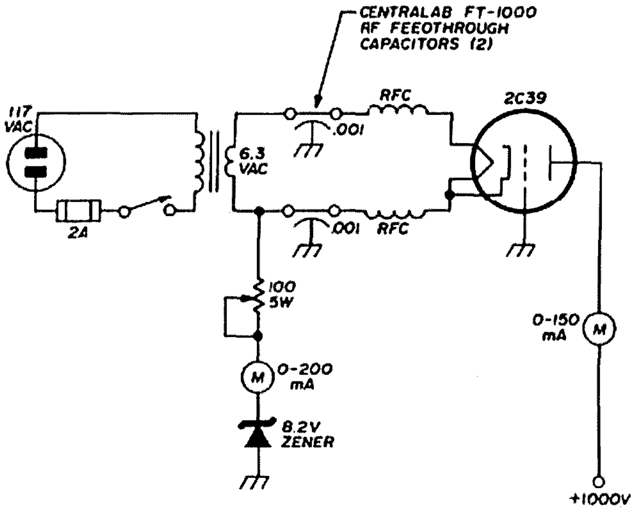


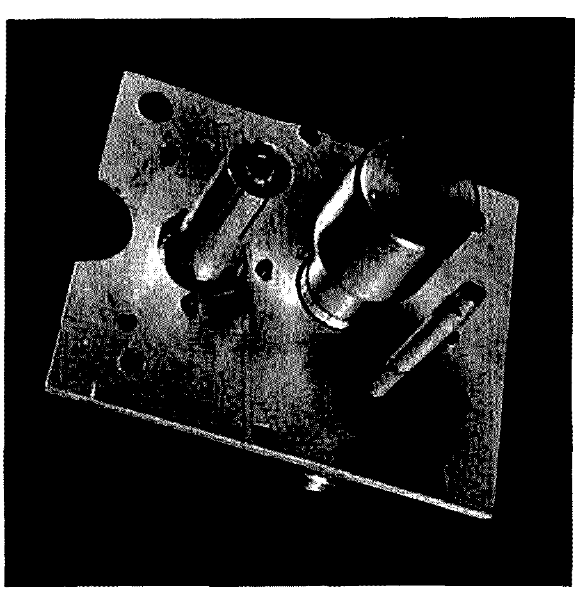
fig. 1. Circuit diagram for the 1152 to 2304 MHz power doubler using a 2C39 which provides approximately 5 dB gain. Rf chokes are 8 turns no. 18 tinned copper, airwound on 3/16" (5mm) mandrel, turns spaced slightly.

been solicited, but none have yet been tried. It is hoped that this article, together with the preceeding one which described the 30-watt power amplifier, will provide the impetus necessary to develop new interest in the 2304 MHz band.

frequency doubler

Having arrived at a reasonably good basic design, attention was focused on

Complete cathode partition assembly showing the heater/cathode line (left), piston tuner (center) and cathode coupler (right).



developing a companion doubler stage capable of driving the power amplifier to full power output with drive to spare. Rather than starting from scratch, it was decided to convert one of the earlier power amplifiers into a doubler by lengthening the cathode cavity to 1-3/8 inch (35mm). No changes were made to the amplifier plate circuit. The resulting doubler circuit is very similar to the power amplifier.

Because the doubler plate circuit is identical to that of the companion power amplifier, only the doubler cathode circuit will be described here. Details of the plate circuit can be obtained from reference 1.

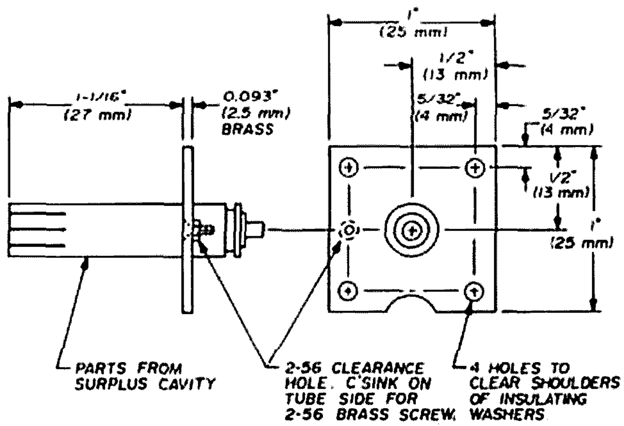


fig. 2. Heater/cathode assembly uses parts from surplus 2C39 amplifier. Use the mounting plate as a template to locate the four 2-56 tapped holes for assembly with the cathode partition.

heater-cathode assembly

The heater-cathode assembly shown in fig. 2 is nearly identical to the one used in the amplifier except that the heater-cathode line extends into the cavity 1-1/16 inches (27mm). The 2-56 brass flat-head machine screw holds a solder lug to connect the cathode side of the heater to the appropriate circuitry (this detail was omitted from the amplifier article). Attach the screw to the plate with a 3/16 inch (4.5mm)

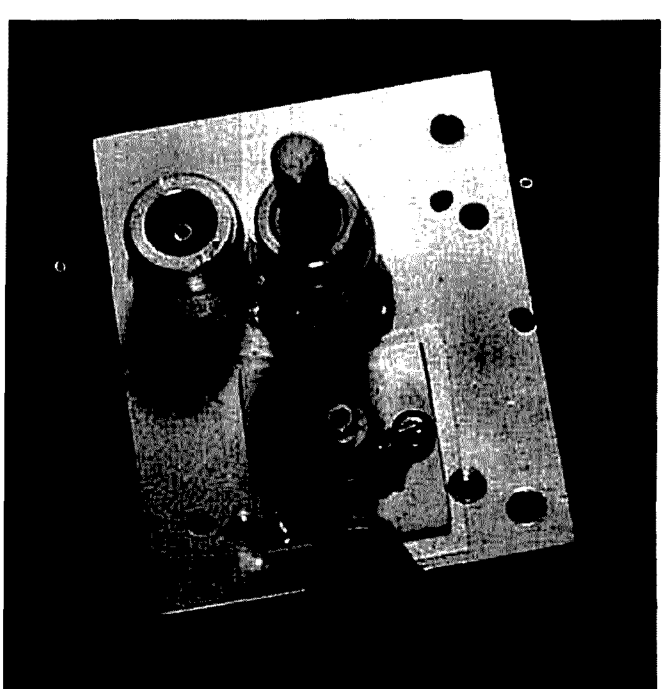
hexagonal nut and then sand the inside surface of the plate flat to make sure the screw head does not project beyond the surface of the plate. Finally, position the plate over the finger-stock assembly and solder the two together as shown in fig. 2.

The completed heater-cathode assembly should be insulated from the cathode partition with insulating shoulder washers and 0.005 inch (0.1mm) Teflon sheet.

cathode piston tuner

The cathode partition is identical to the one described for use with the amplifier and is shown in fig. 3. Rather than using a brass bushing from an old volume control for the piston trimmer, a 3/4 inch (19mm) diameter brass cylinder 5/8 inch (16mm) long is soldered to a length of 1/4 inch (6.5mm) diameter brass rod. Then a 1-1/4 inch (32mm) length of 3/8-32 threaded brass sleeving is slipped over the 1/4 inch (6.5mm) shaft and soldered in place as shown in fig. 4. Soldering should be done with the aid of a propane torch, using solder sparingly. Finally, the tuner and sleeve assembly are screwed into the tuner bushing from inside the cathode partition.

Exterior view of the cathode partition.



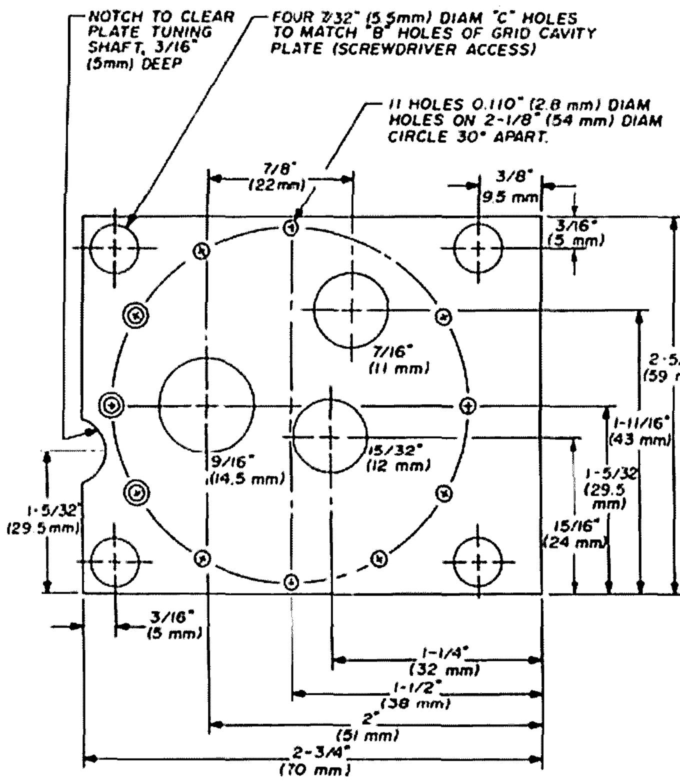


fig. 3. Cathode partition. Holes marked with the letter C are screwdriver clearance holes to facilitate assembly. Material is 0.093" (2.5mm) brass.

If 3/8-32 threaded brass tubing is not available, use 3/8-28 threaded lamp fixture brass tubing which is obtainable in most hardware stores. In this case, the tuning shaft should be made of 5/16 inch (8mm) diameter brass rod to fit inside the lamp hardware. Test the threaded tubing with a magnet to make sure it is not brass-plated steel. Before reassembly with the cathode cavity, the cathode input coupling circuit is assembled as described in the next section.

Helical springs are used to put pressure on the threads of the tuning piston trimmer threads. These springs, which are used on the amplifier as well as the power doubler, are 9/32 inch (7mm) inside diameter. A brass collar with set screws is used at the far end of the tuning shaft to place the spring in compression. These springs are quite important

as tuning is likely to be erratic if they are omitted.

cathode coupling assembly

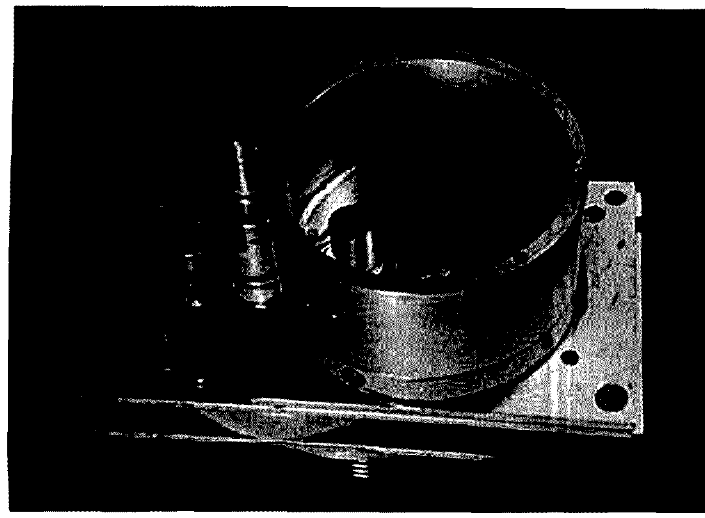
The 1/8 inch (3mm) and 5/32 inch (4mm) OD brass tubing needed to fabricate the cathode coupler (fig. 5) can be obtained from most hobby shops. The connector end of the assembly is soldered to the center conductor of the type-N coaxial connector. The 1/8 inch (3mm) end is slotted with a fine (32 teeth per inch) hacksaw blade. Spread the slotted end slightly to provide a tight slide fit with the cavity end of the coupler assembly.

The cavity end is screwed into the 3/16 inch (5mm) diameter hole in the grid cavity plate. Solder a flat brass washer on the cavity end as shown to provide a good rf contact on the inside of the cavity.

When assembling the cathode partition on the cathode cavity, slide the connector end of the coupler into the cavity end. If the instructions have been carefully followed, the two parts should slide together without interference.

tuning up

The circuit diagram of the frequency doubler shown in fig. 1 is identical to the amplifier wiring diagram except for



View of cathode cavity before installation of the cathode partition. Since an earlier model of the 2304 MHz power amplifier was modified for use as a doubler, parts in this do not correspond exactly with figs. 2 through 5.

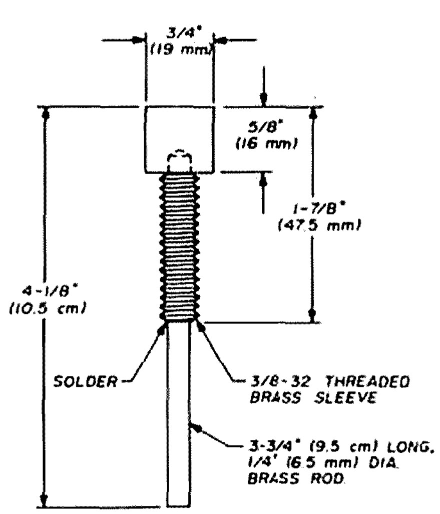


fig. 4. Cathode piston tuner assembly. Brass rod is soldered to 3/8-32 threaded sleeve.

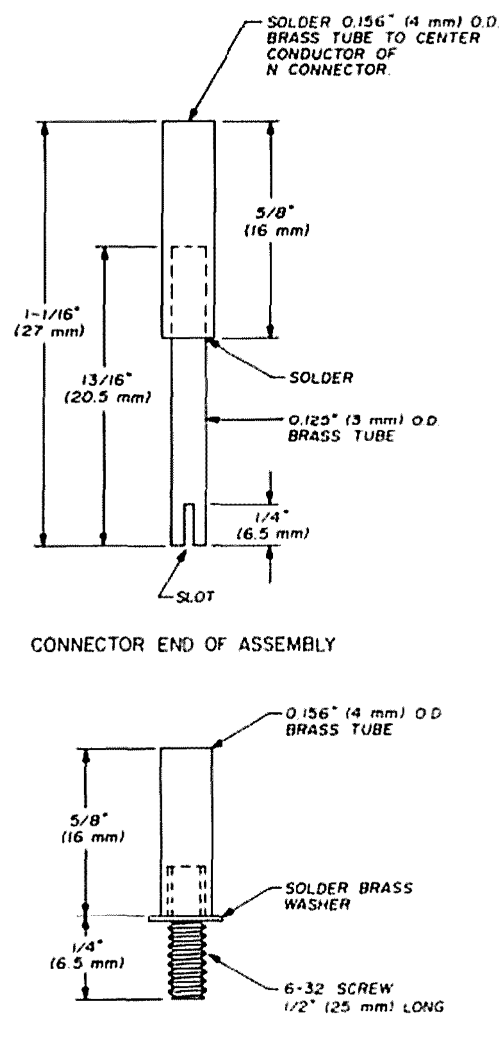
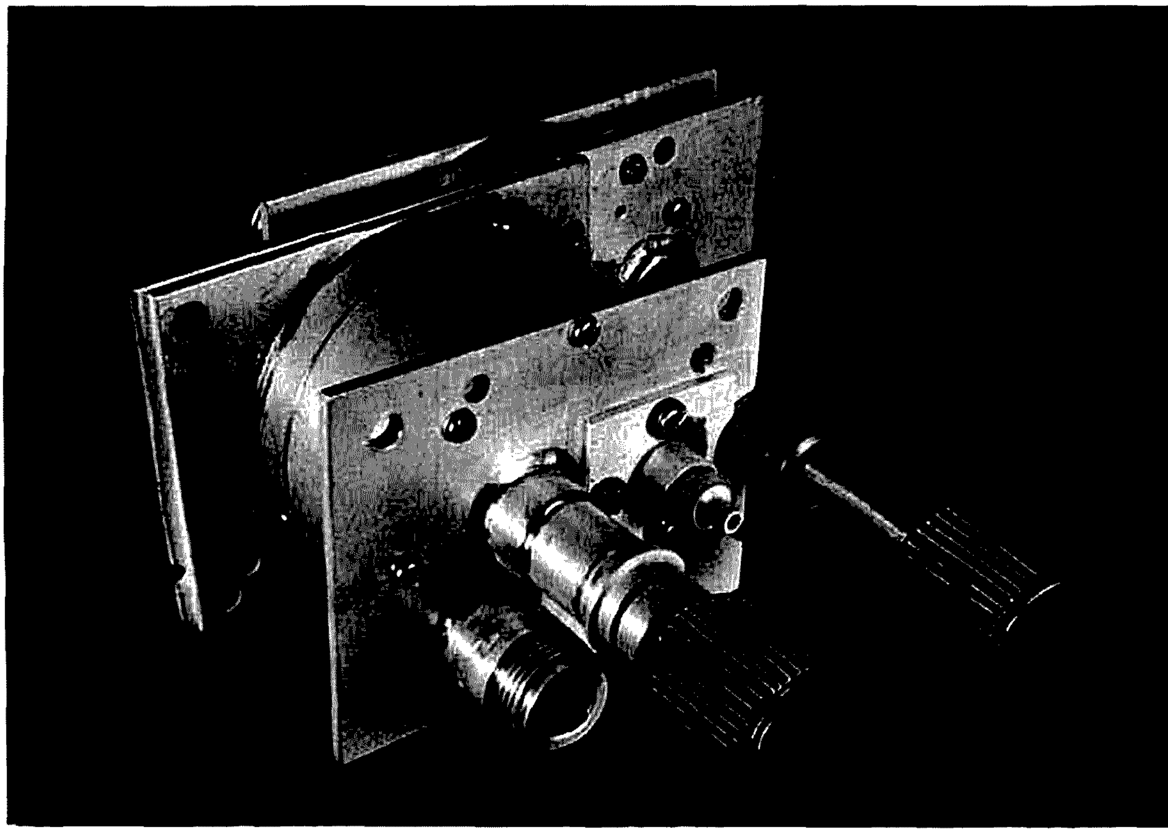


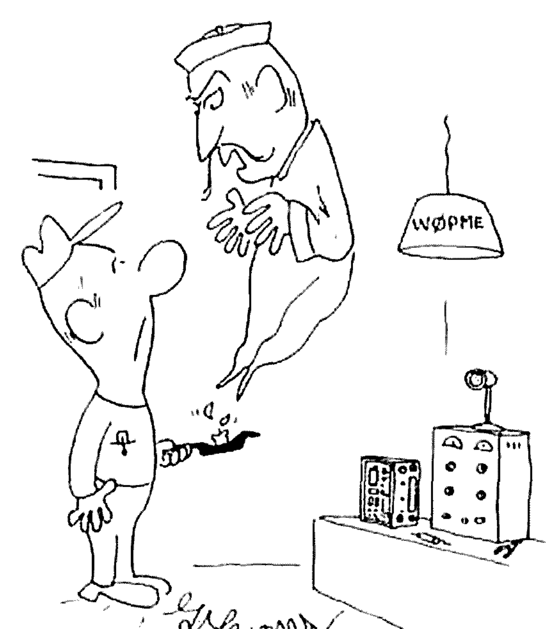
fig. 5. Cathode coupling assembly. Brass tubing can be obtained from most hobby shops.



Front view of completely assembled 1152 to 2304 MHz power doubler.

component values and the meter ranges. A 7.6-volt zener diode is used instead of the 6.3-volt unit to reduce the conduction angle of the plate current for better doubler efficiency. The quiescent (no-drive) plate current should be set to approximately 25 mA by adjusting the position of the slider on the 100 ohm variable resistor.

Measurements were made using a calorimeter which indicate that the doubler has a power gain of approximately 5 dB. It is interesting to note that this doubler provides about 8 dB more power output on 2304 MHz than would be expected from a varactor doubler with the same drive power. Therefore, with less than one watt of drive at 1152 MHz, more than sufficient output is obtained to drive the power amplifier stage to full output. It is recommended that the primary winding of the doubler plate supply transformer be controlled with a variable transformer so that drive to the power amplifier can be adjusted to the desired level.

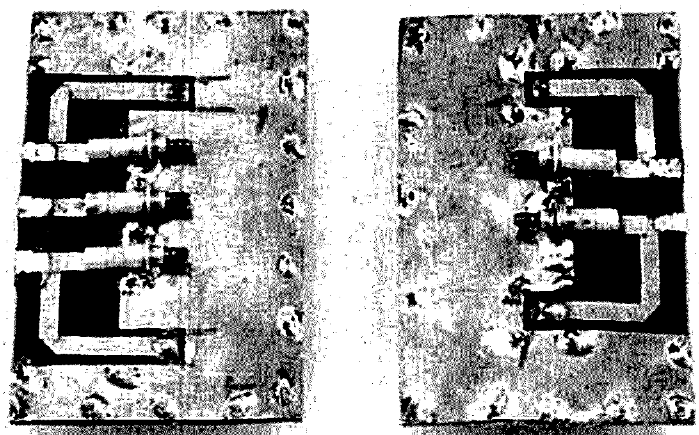


"Just replace capacitor C10 . . . than you'll be back on the air."

reference

1. Norman J. Foot, WA9HUV, "Power Amplifier for 2304 MHz," *ham radio*, February, 1975, page 8.

ham radio



microstripline bandpass filters for 1296 MHz

Miniature bandpass filters for the amateur 1296-MHz band

H. Paul Shuch, WA6UAM, 14908 Sandy Lane, San Jose, California 95124

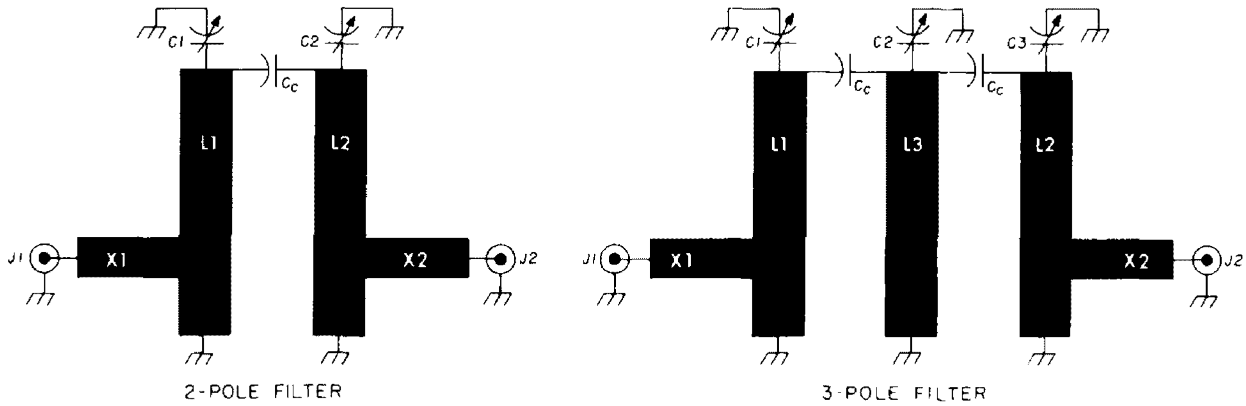
Uhf experimenters frequently need to filter out spurious or image responses, usually with coaxial or trough-line resonators.[1-5] Although properly designed coaxial and trough-line filters offer exceptional skirt selectivity and minimum insertion loss, they are large and bulky and require access to sheet-metal cutting and forming equipment. The 1296-MHz filters presented here are based on printed-circuit microstripline techniques and are easily duplicated in the home workshop.

Two- and three-pole bandpass filters for 1296 MHz are shown schematically in **fig. 1**. In each of the filters parallel-resonant sections, consisting of microstripline inductors and piston trimmer capacitors, are loosely top coupled. The input and output striplines are tapped down on the inductors to provide a match to 50 ohms. The two-pole bandpass filter is functionally equivalent to the filters used at the input of the RF and LO ports of my 1296-MHz double-balanced mixer.[6] In the design presented here, however, the coupling capacitor, C_c, formerly a 0.5 pF chip capacitor, has been replaced by the stray coupling capacitance between the stator ends of trimmers C1 and C2.

As can be seen from the swept frequency response curve in **fig. 2**, these microstripline filters are relatively low-Q devices. The steepness of the rejection skirts may be sacrificed somewhat to minimize passband insertion loss, which for this design averages around 1 dB.

construction

Full-size artwork for the printed-



C1-C3	1-5 pF ceramic piston trimmer	L1, L2, L3	Microstripline inductor, 0.5" (13mm) long, 0.1" (2.5mm) wide, spaced 0.3" (7.5mm) center to center. Bottom ends strapped to groundplane with thin copper strap
C_C	Stray coupling capacitance between stator ends of trimmer capacitors	X1, X2	50 ohm microstripline, 0.1" (2.5mm) wide, any length. Centerline tapped to L1 and L2 0.2" (5mm) from grounded end
J1, J2	SMA or equivalent microstripline launchers (E. F. Johnson 142-0248-001 or similar)		

fig. 1. Two- and three-pole microstripline bandpass filters which tune the range from 1100 to 1500 MHz. Full-size printed-circuit layouts for these filters are shown in fig. 3.

circuit microstripline filters is shown in **fig. 3** and is designed for 1/16 inch (1.5mm) thick G-10 epoxy-glass printed-circuit board, double clad with 1 ounce copper.* The unetched side of the board serves as a groundplane. Board dimensions are such that the filters mount easily in a miniature die-cast aluminum box such as a Pomona 2417. The cutaway view of **fig. 4** shows the method of mounting the piston trimmer capacitors on the circuit board.

With the circuit values shown, these filters can be adjusted to resonate anywhere in the range between 1100 and 1500 MHz. The easiest method to adjust for resonance at 1296 MHz is to connect a weak-signal source through the filter into a receiver, and adjust the trimmer capacitors for maximum received signal. Since the output impedance of the signal source and the input impedance to the receiver may deviate substantially from 50 ohms, it's a good idea to temporarily install fixed attenuators at the input and output of the filter while tuning as shown in **fig. 5**. There is a certain amount of interaction between the trimmer capacitors so the adjust-

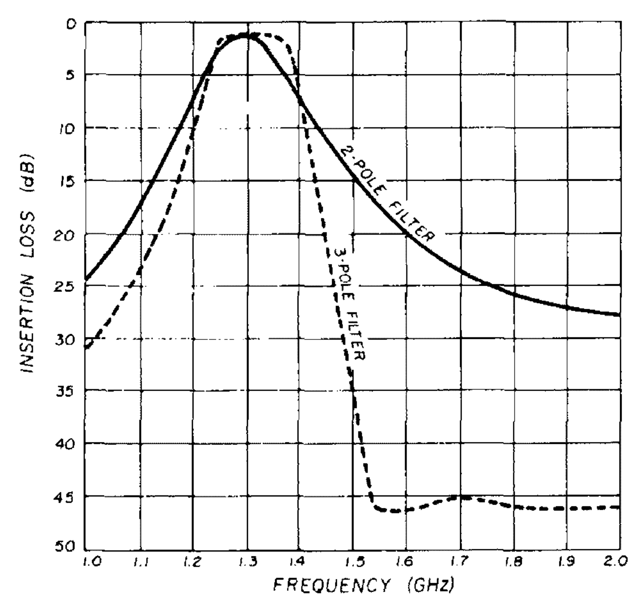


fig. 2. Swept frequency response of the two- and three-pole microstripline filters (measured with a Hewlett-Packard network analyzer and X-Y plotter). The 3 dB bandwidth is 150 MHz and passband insertion loss is about 1 dB. The 20-dB bandwidth is 320 MHz for the 3-pole filter, 570 MHz for the two-pole design.

*Tuned and tested two- and three-pole bandpass filters for 1296 MHz are available from Microcomm. For complete specifications and prices, send a self-addressed, stamped envelope to Microcomm, 14908 Sandy Lane, San Jose, California 95124.

ments should be repeated several times to insure that you have the filters tuned for minimum insertion loss.

If the filter is to be used to reduce the spurious output of a local-oscillator chain, alignment to the desired passband frequency is most easily accomplished by placing the filter in the line between the LO and the mixer and adjusting the filter for maximum indicated mixer current (fig. 6).

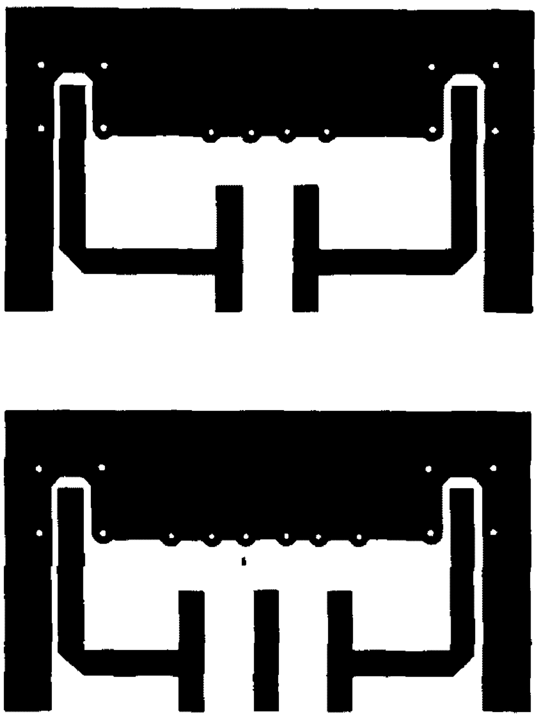


fig. 3. Full-size artwork for the two- and three-pole bandpass filters for 1296 MHz which are designed for 1/16" (1.5mm) double-clad G-10 epoxy-glass circuit board.

applications

Most amateurs who are active on 1296 MHz will probably want to have several of these bandpass filters available on their workbench. In general, accurate measurements on any two-port device are enhanced by the application of filtering at each port. Microstripline amplifiers, for example, tend to be extremely broadband; since transistors tend to have higher gain at lower frequencies, any low-frequency spurious which is applied to the amplifier will be ampli-

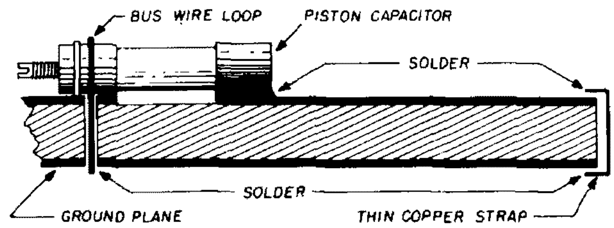


fig. 4. Method of mounting the piston trimmer capacitors on the microstriplines.

fied more than the desired in-band signals. It is not unlikely, in fact, for lower frequency, out-of-band signals to actually force an amplifier into gain compression. Bandpass filters at the input and output of an amplifier under test will thus aid considerably in making accurate gain and dynamic range measurements.

In operational equipment it's a good idea to place bandpass filters between each wideband stage as shown in fig. 7. The filter's 1 dB or so of insertion loss is more than offset by the elimination of image signals and spurious responses. For maximum image rejection it is recommended that the more selective three-pole filter be installed between all active stages. In the local-oscillator chain, where harmonically related spurious signals are separated from the passband by an octave or more, the simpler two-pole resonators are usually sufficient.

acknowledgements

I would like to thank Marvin Wahl, W6FUV, for critiquing the design of these filters, and Stu Rumley, WB6LOU, for assisting in the swept-frequency response measurements.

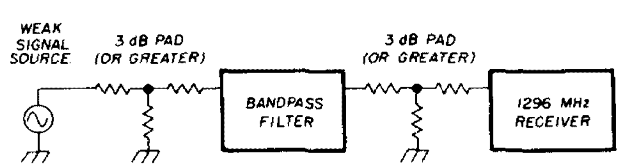


fig. 5. Using a weak-signal source to align a filter to 1296 MHz. The 3 dB attenuators swamp out any impedance mismatches.

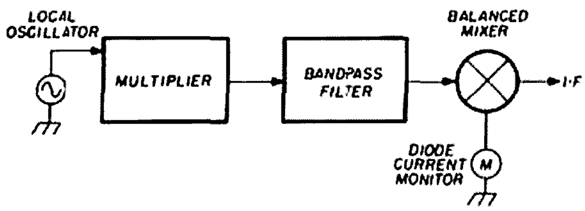


fig. 6. Bandpass filter can be adjusted to the local-oscillator output frequency by tuning the filter for maximum mixer current.

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5. William O. Troetschel, K6UQH, "1296 Revisited," *QST*, July, 1973, page 40.
6. H. Paul Shuch, WA6UAM, "How to Use Double-Balanced Mixers on 1296 MHz," *ham radio*, July, 1975, page 8.

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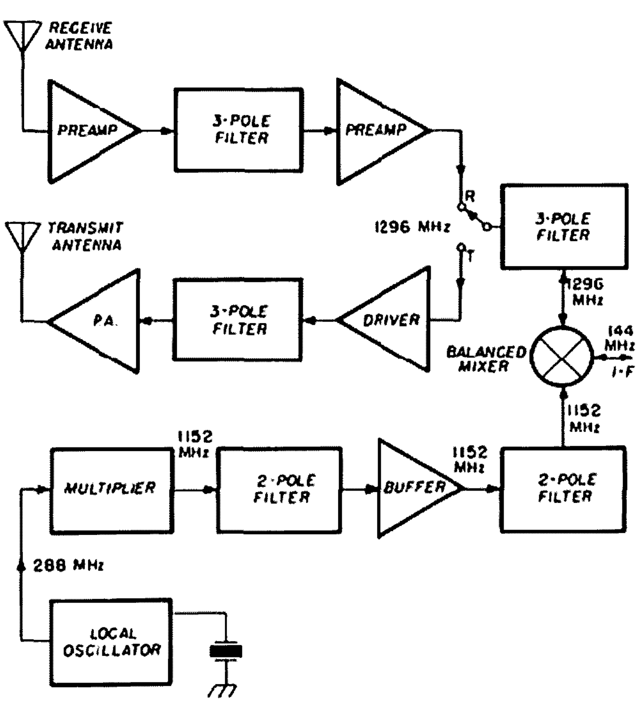

fig. 7. Installation of bandpass filters in a typical 1296-MHz transmitter and receiver. Three-pole filters are recommended between active stages, as discussed in the text.

uhf frequency scaler

New Fairchild 11C90 decade counter IC is a direct plug-in replacement for the popular 95H90 that extends operation to above 500 MHz

Doug Schmieskors, WB9KEY, 22065 McClellan Road, Cupertino, California

Fairchild Semiconductor has introduced another outstanding IC in the 11C00 series[1] which should be of immediate interest to amateurs – the 11C90, a pin-for-pin replacement for the popular 95H90 that has a *minimum* guaranteed toggle frequency of 520 MHz from 0°C to +75°C. At its best, the new 11C90 is a complete front end for a 700 MHz frequency counter (typical toggle frequency at 25°C).*

The 11C90 uhf divide-by-10/11 prescaler makes use of Fairchild's Isoplanar II technology for high speed with reasonable power dissipation. Pins which were unused on the 95H90 decade prescaler are used on the 11C90 to provide a reference voltage which centers the input clock voltage about the switching threshold and allows direct capacitive coupling to the signal source or test antenna. An on-chip ECL-to-TTL level converter is capable of driving ten TTL loads and eliminates the need for any external output interface circuitry.

circuit operation

To take full advantage of the 11C90's uhf counting ability, a circuit such as that shown in **fig. 1** should be built or derived from an existing 95H90 layout.[2,3] Pin 13 (TTL V_{EE}) should be tied to ground (low) if the TTL output (pin 11) is used. If only the ECL output

*The 11C90 is available now from franchised Fairchild distributors worldwide for $16.00 in small quantities.

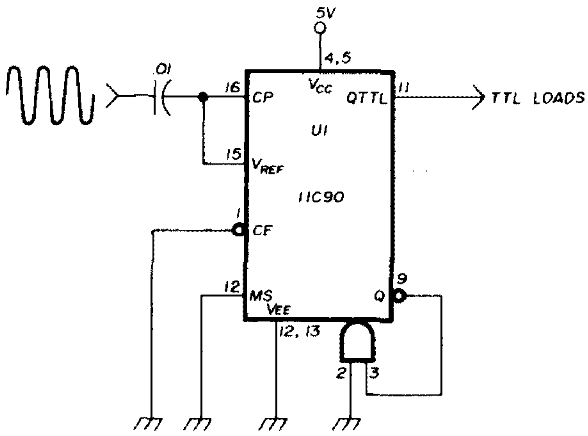

fig. 1. Divide-by-10 uhf prescaler has a minimum guaranteed toggle frequency of 520 MHz. Typical toggle frequency at 25°C is 700 MHz.

(pin 8) is used, pin 13 may be left open to reduce power consumption.

A reference voltage is generated internally across a 400-ohm resistor to the V_{BB} supply and is present at pin 15 (V_{ref}). This completely eliminates any need for an external biasing network.

Pins 6 and 7 of the 11C90 are uncommitted 2000-ohm resistors which are internally connected to the mode control inputs, $\overline{M1}$ and $\overline{M2}$. When tied high (+5 volts) these resistors allow the associated mode control input to be driven from TTL; if these inputs are left open or tied low, the mode control inputs offer, respectively, unterminated or terminated ECL loads to the drivers.

The mode control inputs are useful primarily when the 11C90 is employed in the divide by 10/11 mode to produce non-standard divide ratios such as those used in pulse swallowing for frequency synthesis. The 11C90 logic symbol (fig. 2) and truth table in table 1 should aid in understanding the device.

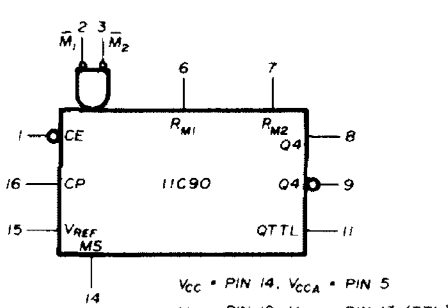

fig. 2. Logic symbol for the Fairchild 11C90. Mode control inputs $\overline{M1}$ and $\overline{M2}$, and RM1 and RM2 inputs are discussed in text. The IC includes a built-in ECL-to-TTL converter.

Circuit layout, although not critical, can be used to enhance the high-frequency operation of the 11C90. Proper power supply decoupling, broad ground connections, short signal runs, and short leads (sockets are *not* recommended) will all help the user to reap the maximum performance that has been built into the device. The 11C90 typically requires only 65 mA as compared to 90 mA for the 95H90, so it runs much cooler than its predecessor.

table 1. Mode selection for the 11C90. Low is indicated by L, high by H.

M1	M2	module divide by
L	L	11
H	L	10
L	H	10
H	H	10

summary

The 11C00 family of sub-nanosecond logic now consists of nine devices ranging from the 11C05 prescaler to the 11C01 gate package, and includes the 11C58, a 150-MHz voltage-controlled monostable oscillator which features a 4:1 frequency range with 2-volt dynamic range. These new devices obviously open up a whole new range of frequency synthesizer possibilities, but that's another story.

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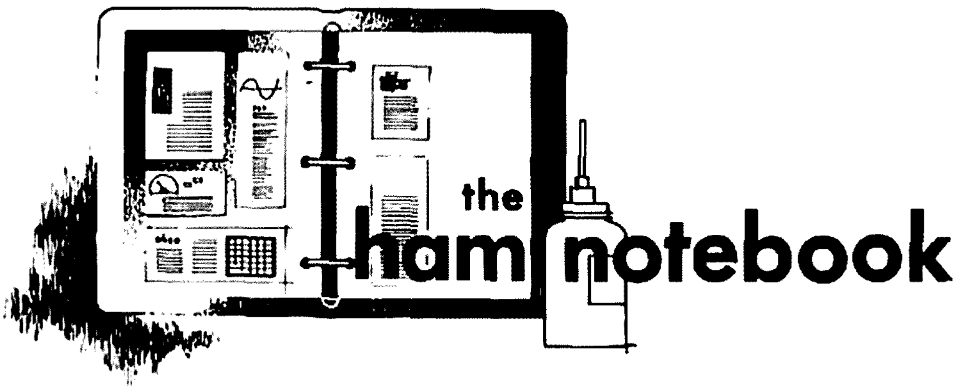
2. F. Everett Emerson, W6PBC, "300 MHz Divide-by-Ten Frequency Scaler," *ham radio*, September, 1972, page 41 (correction, December, 1972, page 90).
3. F. Everett Emerson, W6PBC, "Circuit Improvements for the Advanced Frequency Scaler," *ham radio*, October, 1973, page 30.

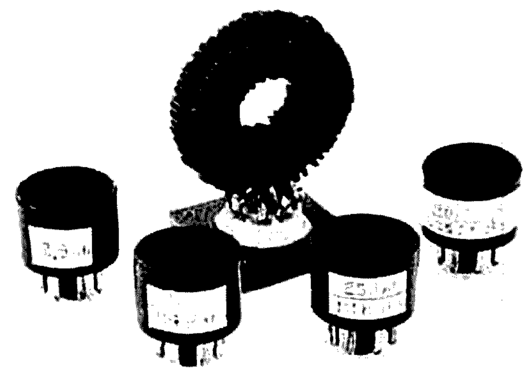
ham radio



quadrifilar toroid

The prevalence of roller inductors in transmitters and antenna couplers attests to the need for adjustability. The quadrifilar toroid limits the adjustability to discrete steps but offers the advantages of small size, internal field requiring little if any shielding, and balun applications.

The ends of the four, parallel, tightly-coupled windings are connected into the desired configuration by an octal socket and tube base. Of several octal sockets I tested those with "wrap-around" pins consistently measured



Construction of the quadrifilar toroid which is based on an Amidon T-200-6 powered-iron toroidal core.

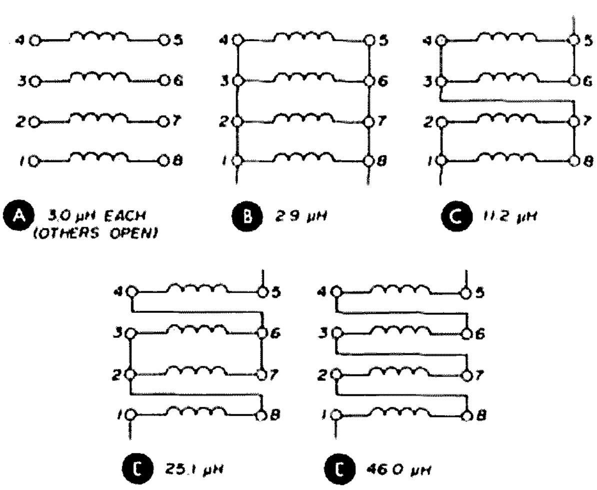


fig. 1. Five methods of interconnecting the four quadrifilar windings, and the measured inductance values of *each* configuration. Winding consists of 16 quadrifilar turns of no. 12 (2.1mm) on an Amidon T-200-6 toroid core.

0.003 dc ohm per contact while the "edge-bite" pins varied from 0.003 ohm for a few pins to many times that value for most. Obviously only the "wrap-around" octal socket is recommended and preferably in ceramic or mica-filled bakelite. If a *low-resistance* 12-point switch or plug-socket can be found, a hexifilar toroid with inductance ratios of 1, 4, 9, 16, 25 and 36 can be built.

This example of a quadrifilar toroid consists of an Amidon T-200-6 toroidal core with four windings of 16 turns each of number 12 (2.1mm) enamelled

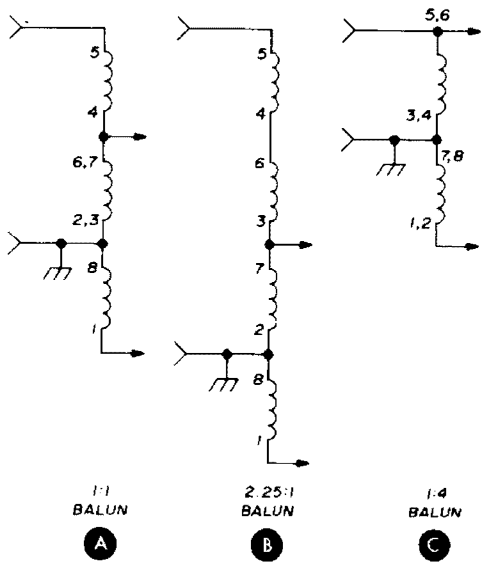


fig. 2. Quadrifilar windings may also be used to build baluns with 1:1, 2.25:1 and 1:4 transformation ratios.

wire. **Fig. 1** shows the socket connections and the measured inductance values.

In balun service the four independent windings lend themselves to several configurations; three of the simpler forms are shown in **fig. 2**. The easy access to the terminals suggests other arrangements.

The frequency response of the 1:1 balun is flat to at least 20 MHz (the limit of my sweep generator) and probably well beyond. Even at one-half and twice termination the smooth roll-off dropped only 30 and 20 per cent, respectively, at 20 MHz.

R.S. Naslund, W9LL

technique speeds antenna tuner adjustment

This article deals with a simple and accurate procedure for tuning or adjusting antenna tuners without using a transmitter *or* a standing-wave bridge. To avoid some possible confusion the term antenna tuner refers to such devices as Johnson Matchbox, Millen Transmatch, Murch Ultimate Transmatch and most similar homebuilt antenna tuners. Every ham shack should have at least one.[1]

In almost every technical article on antenna tuners that is published, you are instructed to make a written record of the dial settings and coil tap points for future use. If you've gone through this you know it's time consuming to search for coil tap points and tune two or three variable capacitors for each tap, trying to find the correct settings for each operating frequency. Furthermore, going through this procedure on the air generates a lot of unnecessary interference. I deliberately put up a four-band parallel dipole so I could avoid using (or adjusting) an antenna tuner, but because of the high swr the antenna tuner is now back in the line.

The simple technique discussed here for adjusting your antenna tuner does require an additional piece of test equipment which you may not have. However, the necessary test gear, a simple impedance bridge, can be easily built from junkbox parts. Although several RX impedance bridges have been described in the amateur literature,[2,3,4] the more simple *antennascope*[5] or antenna impedance meter[6] are suitable for

1. Ed Noll, W3FQJ, "Antenna Tuners," *ham radio*, December, 1972, page 58.
2. Henry S. Keen, W2CTK, "A Simple RX Bridge for Antenna Measurements," *ham radio*, September, 1970, page 34.
3. Bill Wildenhein, W8YFB, "A Low Cost RX Impedance Bridge," *ham radio*, May, 1973, page 6.
4. Jerry Hall, K1PLP and John Kaufmann, WA1CQW, "The Macromatcher," *QST*, January, 1972, page 14.
5. William Orr, W6SAI, *Beam Antenna Handbook*, Radio Publications, Wilton, Connecticut, page 178.
6. Robert G. Middleton, *101 Ways to Use Your Ham Test Equipment*, Howard Sams & Company, Indianapolis, page 80.

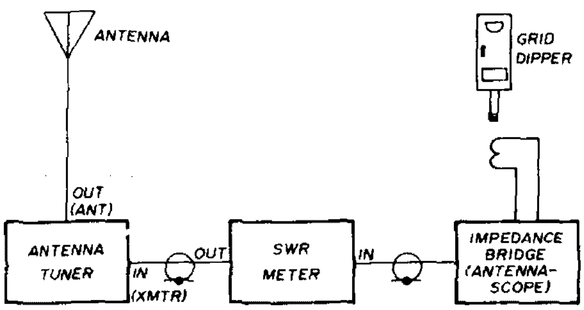


fig. 3. Simple test setup speeds initial adjustment of antenna tuning unit with minimum on-the-air interference.

this application. You will also need a grid-dip meter or low power transmitter as a source of rf for the impedance bridge. A grid-dip meter is highly recommended as it will cause less unnecessary interference.

impedance bridge. When finding the tap point it is suggested that the clip be held by its insulation and moved slowly up and down the coil until you see a downward movement of the bridge meter. That's the tap point you're looking for. This procedure is simplified somewhat if your antenna tuner uses a roller inductor, but the end result in either case is the same.

When the correct tap point has been found, fasten the clip on the inductor and tune the variable capacitor for as perfect null as possible on the bridge meter. Record the dial settings for future use. When a transmitter, tuned to the same frequency, is connected in place of the impedance bridge, only very minor touchup of the antenna tuner should be required for an indicated vswr of 1:1. The data of table 1 show the results I obtained while using this procedure to adjust a Transmatch.[7] Note the close correlation between dial settings obtained with three different types of impedance bridges. The operating swr for all cases was very nearly 1:1.

table 1. Comparison of Transmatch dial settings obtained with three different impedance bridges using the test setup of fig. 3.

bridge type	frequency (MHz)	input capacitor	inductor tap	output capacitor
RX Bridge[3]	3.95	98	59	40
	7.25	90	68	0
	14.05	55	3	40
	21.05	90	2	10
Macromatcher[4]	3.95	95	60	15
	7.25	100	68	0
	14.05	35	4	24
	21.05	90	2	15
Antennascope[5]	3.95	90	52	10
	7.25	98	68	0
	14.05	30	4	20
	21.05	95	2	15

Set up the test equipment as shown in fig. 3. If you use a grid-dip meter you won't get a reading on the swr meter, but at this point that's not important. Set the impedance bridge to 50 ohms (or 75 ohms if that's the impedance of your transmission line), tune the grid-dipper to the desired operating frequency and couple it to the impedance bridge. The meter on the bridge should swing upscale.

Now locate the tap on the antenna tuner inductor that causes a null on the impedance bridge.

Howard Stark, WA4MTH

7. Lewis G. McCoy, W1ICP, "The Ultimate Transmatch," *QST*, July, 1970, page 24.

burglar-proof alarm

When you are setting up a burglar-proof alarm for your car, you should have an unusual alarm. The more unusual the alarm, the harder it is for a burglar to get into the car. The most important parts of the alarm are the switches used to activate it. These switches must be placed so that they are hard to find, but still allow complete protection. This means that switches should be used to prevent the car from being towed away, as well as being broken into.

After the switches have been placed, you must connect them to some sort of alarm. The alarm device must make a very noticeable sound. This requirement rules out the car's horn because people hear them constantly in a populated area. The best device for the alarm is a siren. There are two types of sirens that can be used, mechanical or electronic. Both types are suitable for the system shown in **fig. 4**.

In the schematic there is a time-delay switch which is used to eliminate outside control on the car. This is important because it gives the advantage of surprise when a burglary is being committed. The approximate one-minute time delay allows you to enter the car and shut the alarm off. This is enough time to shut the device off if you know how, but not enough if you don't. This stops the burglar from removing anything that is fastened inside the car or searching the interior. The schematic also shows that two switches are used in the driver's door. The second switch is used to activate the circuit with the time delay.

The on-off switch is a simple dpst switch placed somewhere in the middle of your ham gear. This way a burglar will never realize that it is the switch to deactivate the device. Also, the battery and siren are placed in the trunk. This makes it very hard for a burglar to disarm the system.

As noted in the schematic, there are four other switches marked "trunk," "hood" and "limit." These four switches are very important in deterring a person from stealing your car. The trunk and hood switches are simply placed in the trunk and hood, preventing anyone from opening either one and tampering with anything inside.

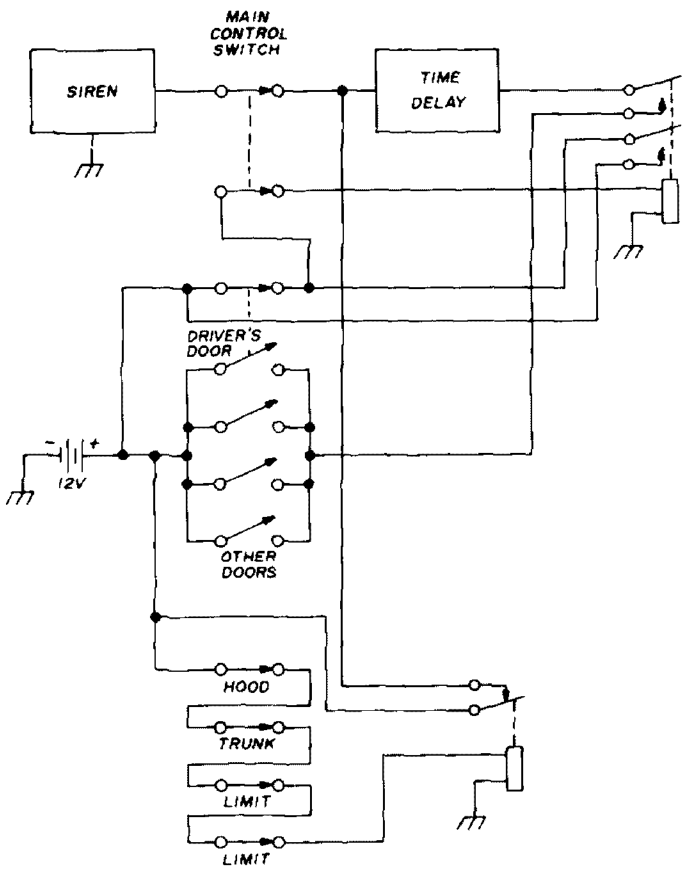


fig. 4. Simple burglar-proof auto alarm. All switches are shown as they would be when the doors, hood and trunk are closed.

The other two switches are harder to place, but they prevent the car from being moved. One of the switches is placed inside the car and is operated by the parking brake cable. Cutting the cable releases the tension on the cable and activates the alarm. The other switch is placed on one of the back shock absorbers. It is a limit switch, operating when the shock absorber is extended to its maximum. This sounds the alarm if the car is being towed away.

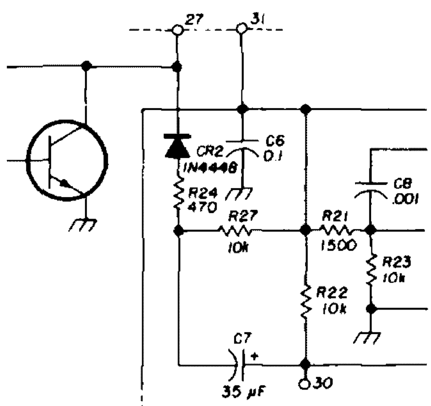
Installed correctly, this alarm system will keep your car well protected. It has already prevented three burglaries for me.

Glenn Eisenbrandt, Jr.

short circuits

universal tone encoder

A few transceiver models using diode PTT switching will not operate correctly with the Universal Tone Encoder shown in the July, 1975, issue. The problem occurs in the tone-burst mode after the PTT button is released, and results from the charging current drawn by C7. Some transceivers are not able to supply this current and will not return fully to the receive mode.



This problem is solved by adding CR2 and R27 as shown in the schematic above; the polarity of C7 was also reversed in the original schematic. The encoder will now operate with relay- and diode-switched transceivers. The new circuit board incorporates this change. Circuit boards are available from Larry McDavid, W6FUB, 185 South Alice Way, Anaheim, California 92806.

dc latch circuit

In the CMOS dc latch circuit, fig. 2, on page 44 of the August, 1975 issue, the D input of U2A (pin 5) should be connected to the $\bar{Q}$ output (pin 2), *not* to the Q output (pin 1) as shown.

low-frequency loop antenna

In the article on the loop antenna receiving aid in the May, 1975 issue, no ground return is shown for the fet preamp (fig. 3) or Q multiplier (fig. 4). In both cases the 100k resistor connected to the gate of the HEP802 fet should be grounded.

automatic az/el control

Several errors appeared in the automatic azimuth/elevation rotator control system published in the January, 1975, issue of *ham radio*. In the base diagram for the 558 op amp (fig. 3), the inverting and non-inverting inputs to the lower op-amp are reversed (pin 5 should go to the non-inverting [+] input). In fig. 7 the *sensed position* output should be connected to the junction of the 100- and 750-ohm resistors, *not* to the op-amp output terminal. Also, add two to all IC numbers in the second column on page 29 and the first column on page 31 (U11C, for example, should be U13C).

Some readers have found that the frequency-selective amplifier in fig. 4 oscillates. This can be easily solved by increasing the value of the shunt resistance of the bridged-T network. In amplifier U1, for example, the resistor to change is the 7500-ohm unit connected between the two 0.05 μF capacitors.

To eliminate difficulty with rf interference, shunt each rotator motor winding lead with a 0.01 μF disc capacitor at the control unit case. Treat the leads to the rotator potentiometer as illustrated in fig. 1, shown here. In addition, the grounded end of the rotator potentiometer should be fastened to the circuit ground near the comparator, U11. Otherwise, unrelated ground currents may upset the sensed rotator position.

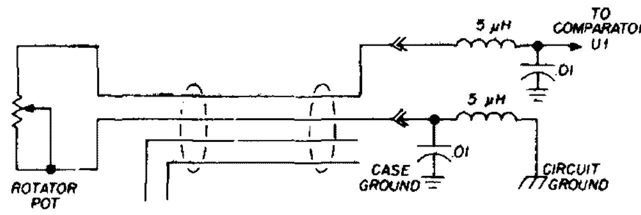


fig. 1. Use the circuit shown here with rotator potentiometer leads to eliminate difficulty with rf interference.

phase modulation techniques

In the article on phase modulation principles and techniques on page 28 of the July, 1975 issue, the value of R in fig. 7 should be 10k. The loss formula shown in fig. 7 should be

$$\text{loss} = 20 \log \left(\frac{X_C}{\sqrt{R^2 + X_C{}^2}} \right) \text{ dB}$$

communications receiver

In the communications receiver described in the October, 1975 issue (page 32), the KVG XF9E crystal filter has 12 kHz bandwidth, *not* 2.4 kHz. Transformers T1 (fig. 2) and T1, T2 and T3 (fig. 8) are wideband transformers which I5TDJ wound on Ferroxcube 0.25" (6.5mm) diameter toroid cores (permeability of 1000 or more). The Amidon T50-6 cores specified in the article have low permeability and low-frequency performance is poor. For those who have asked, Q6 in fig. 8 which is specified as an HEP S0014 may be replaced with a 2N3866 or 2N4427.

Designer I5TDJ has heard from several amateurs who have built duplicates of this receiver that they sometimes have trouble with the fet crystal oscillator circuits (fig. 8). He subsequently tested a number of fets and found that some circuits would not oscillate with fets with high I_{DSS} because of the large voltage drop across the 1000-ohm drain resistor which biased the fet into the pinch-off region. This can be solved by using low I_{DSS} fets or by reducing the value of the drain resistor to 100 ohms.

radiation hazards

In the September editorial W1DTY made an error when calculating the power density at 10 watts input to a 30-foot dish. Since the 10 watts is essentially spread over the area of the dish in the near field (within one or two dish diameters), the power density at 10 watts input is 0.061 mW/cm². An input of 1642 watts would be required to reach 10 mW/cm².

new products

logarithmic speech processor

New from MFJ Enterprises is the LSP-520BX speech processor, which provides 400% more rf power to your phone signal. The LSP-520BX adds this extra punch by means of three active filters, two of which are switch-selectable, and a low-distortion IC logarithmic amplifier with a 30-dB dynamic range, assuring constant transmitter output without clipping or appreciable distortion. Voice frequencies are tailored to put communication intelligence where it will do the most good – in your transmitted signal.

Four techniques are used to maximize the voice-to-noise power *ratio*. First, an rf-protected preamp is optimized for low-noise performance by using a premium, low-noise transistor. Second, putting two high-pass active filters *before* the log amplifier input ensures a clean, noise-free signal at the log amp output. Third, battery operation eliminates hum, and fourth, a filtered and shielded input circuit provides immunity from rf fields.

The low-noise preamp has a gain of about 43 dB. An emitter follower matches the preamp output to a 500-Hz two-pole active filter, which has a rolloff of 12 dB/octave (at 250 Hz the signal is attenuated 12 dB). A switch-selectable two-pole 1400-Hz highpass active filter, which also has a 12-dB/octave rolloff, follows the 500-Hz filter. When these two filters are cascaded, rolloff is 24 dB/octave below 500 Hz for maximum filtering. Following the two filters are a compression-level control and the logarithmic amplifier. A six-pole, low-pass active filter accepts the log amp output. This filter features steep rolloff at 36 dB/octave, with a 2100-Hz cutoff frequency. Thus, bandwidth restriction prevents a *wide* ssb signal and removes distortion products.

Installation is simple. Plug your microphone into the processor, plug the processor output into your transmitter microphone output, and you're ready for some pleasant surprises on the crowded phone bands. The LSP-520BX is priced at $49.95 or $35.95 in kit form. Write MFJ Enterprises, P.O. Box 494, Mississippi State, Mississippi 39762, or use *check-off* on page 142.

525-MHz uhf prescaler

The new Pagel model 525 uhf prescaler divides frequency by ten to extend the range of any 50 MHz or higher counter to the vhf and uhf bands. The unit also contains a 20 dB preamp for the unscaled 1 MHz to 50 MHz range to improve frequency counter sensitivity to 5 millivolts rms or better. Sensitivity is 50 mV rms at 500 MHz, and 30 mV rms below 400 MHz. A through-line

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feature with an internal signal sampler can be used with transmitters up to 100 watts (requires 50-ohm dummy load). This feature can be used to perform simultaneous power and frequency measurements and is a great time saver.

The model 525 operates from the 117 Vac line or battery power (8 to 15 volts) may be used for portable or mobile use. Price is $159. For more information, write to Pagel Electronics, 6742-C Tampa Avenue, Reseda, California 91335, or use *check-off* on page 142.

hamtronics catalog

Hamtronics, Inc., long known for its vhf preamplifiers and fm communications receiver kits for amateur and monitor applications, recently announced a new catalog, which is available to readers in return for a self-addressed, stamped envelope. It lists many new products, including a high performance version of its famous standard vhf preamp. This kit, which is wired in series with the coaxial antenna lead of vhf communications receivers of various operating frequencies, boosts the receive signal by 20 dB or more, depending on the frequency. It operates from +12 Vdc, and is constructed on a PC board. Cost of the kit is $9 (wired and tested, $14).

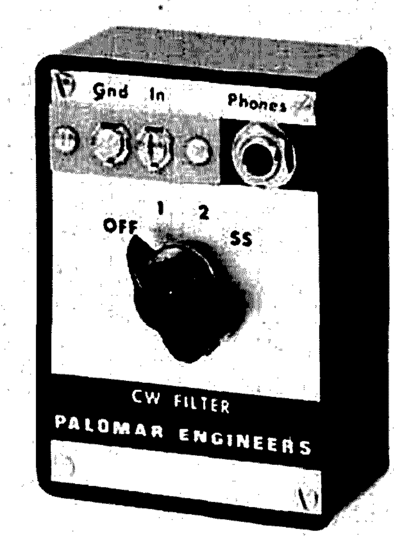
The second new product is a two-stage grounded-gate preamplifier for uhf receivers in the 400-500 MHz range, including amateur, commercial, and monitor receivers. It provides 20 dB gain, and is priced at $15 (kit) or $30 (wired and tested). A companion uhf converter kit is available for operation on various i-f frequencies, thereby converting a vhf receiver into a uhf receiver. The converter kit is priced at $20 plus crystal.

A new improved vhf receiver for fm communications has also been introduced in this catalog. It consists of a vhf converter board and a i-f/audio board. The converter is also available separately

to operate with various i-f frequencies. The vhf receiver kit is $65 less crystal. A uhf version, which includes the uhf and vhf converters and i-f/audio board, is priced at $80, less crystals.

For your free copy of this new catalog, send a self-addressed, stamped envelope to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

CW filter



Operators active in contest and DX work will welcome a novel CW filter offered by Palomar Engineers, The circuit combines a wideband and narrowband filter to provide simulated stereo reception. Active filters prevent annoying ringing and give sharp skirt selectivity, which removes all signals except those within an 80-Hz bandwidth. The simulated stereo technique allows off-frequency signals to be heard, but because of the action of mind and ears, the off-frequency signals do not interfere with the desired signal.

The filter connects between your receiver and a set of stereo headphones. In the simulated-stereo mode, the narrowband signal is applied to one side of the stereo headset and the wideband signal to the other. Alternatively, the narrow-

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phones by panel-switch selection. The simulated stereo mode uses both filters with a dramatic improvement over either filter alone. The desired signal is heard in both phones; the off-frequency signals and noise are heard in only one phone. The mind concentrates on the desired signal and rejects the interference — yet off-frequency calls can still be heard, which otherwise might be missed.

The center frequency of the CW filter is 800 Hz. Bandwidths of the narrow- and wide-band filters are 80 and 300 Hz respectively. A 9-volt transistor battery supplies power. Input impedance is 1 megohm; the output will drive either low- or high-impedance headphones. The panel switch has four positions: *off* (receiver output direct to phones); wideband amplifier to both phones; narrowband amplifier to both phones; and simulated stereo. The CW filter is $39.95 postpaid in U.S. and Canada. More information may be obtained from Palomar Engineers, Box 455, Escondido, California 92025, or use *check-off* on page 142.

unique ic op-amp applications

A specialist in IC operational amplifiers, Walter Jung, has written this book on the uses of unique op amps. Unique op amps are those with characteristics that set them apart from previous amplifiers. Modified types of op amps are discussed along with totally unique types, such as programmable op amps, operational transconductance amplifiers, and quad current-differencing amplifiers. The material has been extracted from another Sams book, *IC Op-Amp Cookbook*. Heavily illustrated. 144 pages, softbound, $4.95 from Ham Radio Books, Greenville, New Hampshire 03048.

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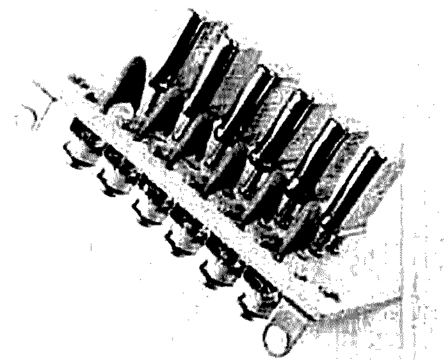
Regency crystal deck



Topeka FM Communications has just introduced a new improved version of their highly popular 6T-HR2 six-channel crystal deck. Mounted in a Regency HR-2 or HR2-A, the deck allows full 12-channel transmit/receive capability. Improvements include a smaller board which makes installation notably easier. Component placement has been changed to allow netting without removing the speaker from the radio.

The new version, designated the 6T-HR2-3, is now available for $15.50. Kit versions are also available for $11.50. For more information, write or call Topeka FM Communications, Inc., 125 Jackson, Topeka, Kansas, 66603.

noise in electronics

Why is noise important? What is shot noise? How can noise figure be measured using a signal generator? These and dozens of other pertinent questions are answered in this new book for the amateur, engineer and technician. They provide the reader with a basic understanding of noise characteristics and noise measurement techniques for practical applications.

The author first introduces the reader to noise with explanations on white, pink, man-made, atmospheric and galactic noise. The remainder of the book answers questions about thermal noise, shot noise, noise bandwidth, special considerations for noise, signal-

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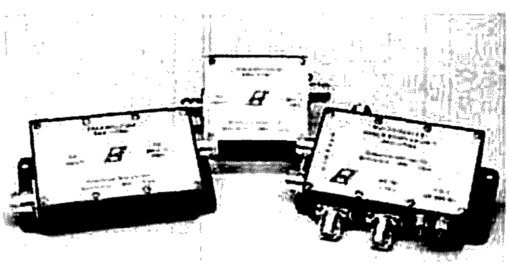
to-noise ratio, noise figure and other miscellaneous noise characteristics.

Here's an opportunity to learn about flicker noise, noise power, the effect a-m detectors have on noise and dozens of other noise-related subjects. Easy-to-understand answers are detailed without the complex mathematical manipulations usually required with noise associated calculations. Illustrations, examples, and tables of solutions are provided to further explain the answers. 96 pages, softbound, $3.95 from HR Books, Greenville, New Hampshire 03048.

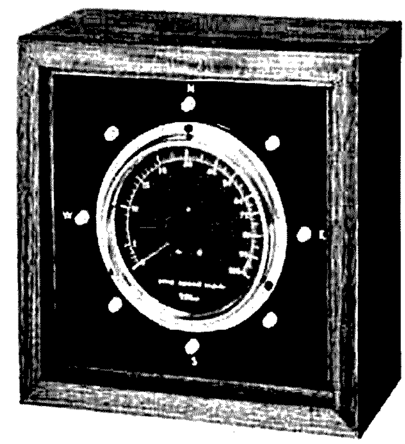
multicoupler/ preamplifier



The new multicoupler set from Radiation Devices features an antenna-located preamplifier and provides preamplification of signals at the base of a broadband high-frequency antenna to overcome coaxial cable loss. Preamplifier BBA-1/PMS-3 has greater than 9 dB gain over the band from 2 to 50 MHz. It receives power via the coaxial cable connecting it to the Multicoupler/Power Adapter Unit MPU-1. The MPU-1 provides four isolated signal ports to receivers or other equipment. Intermodulation and cross-modulation distortion products are greater than 60 dB below the desired signal at zero dBm output level. The unit operates from 115 Vac, 50 to 400 Hz.

For more information, contact Radiation Devices Company, Post Office Box 8450, Baltimore, Maryland 21234, or use *check-off* on page 142.

wind direction and velocity meter



Knowing wind speed and direction will allow you to trim that big beam antenna for minimum wind resistance when the next storm arrives. The model 75C Brunswick Wind Set by TMAC Products consists of a wind-speed transmitter, wind-direction transmitter, an indicator mounted in a handsome console, and all cables and mounting hardware. The transmitter units are low profile; the entire assembly measures only 12-3/4 inches high by 24 inches long (53 by 61cm) and may be mounted on any convenient surface.

The wind-speed transmitter consists of a dc generator coupled to a 5½-inch (14cm) diameter, spherical cup rotor assembly mounted on a 1-inch (2.5cm) diameter pvc pipe support. Wind speed is indicated by a 6-inch (15cm) diameter, 250-degree linear taut band pivot and jewel movement. Readout is in mph, with 1-mph divisions between 0 and 100 mph inscribed in white against a contrasting background.

The wind-direction transmitter uses hermetically sealed reed switches actuated by a magnet in an environment-protected, low-friction assembly. Wind direction is indicated by eight panel lamps, one at each cardinal compass point, located around the periphery of the wind-speed indicator. Intercardinal

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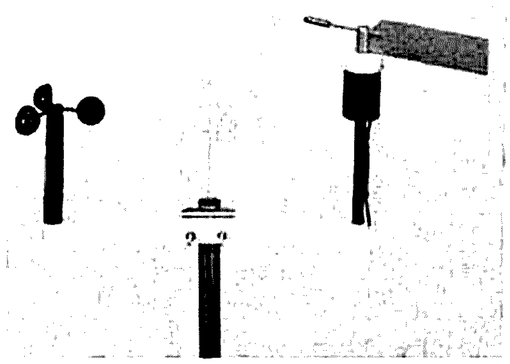


compass points are indicated by the illumination of two adjacent lamps. Thus, 16 compass points may be indicated; at least one indicator lamp will be on at all times. The instrument is powered by 110 V, 60-Hz. Price and additional information are available from TMAC Products, P.O. Box 28341 (Lincoln Village Branch), Columbus, Ohio 43228, or use *check-off* on page 142.

miniature touch-tone encoders

Data Signal has announced a new line of solid-state crystal-controlled Touch-Tone encoders which use a CMOS encoder IC. Only ¼ inch (6.5mm) thick, these self-contained units provide Touch-Tone capability to repeater stations or provide data entry. They are designed to be mounted directly on the side of hand-held portables, on the front of mobile transceivers, or on the dashboard of vehicles. The circuitry is completely rf proof, and all electronics are contained *within* the keyboard. Keyboards with 12 Touch-Tone digits are available in three sizes: 2¼x3 inches (57x76mm), 1½x2 inches (38x51mm) and 2x1½ inches (51x38mm). The 16-digit keyboard is 2 inches (51mm) square. These keyboard encoders, type DTM, require only three external connections and are priced at $49.95.

Also available from Data Signal is a sub-miniature Touch-Tone encoder and keyboard which is designed for use with

hand-held fm transceivers. The encoder PC board measures a mere 0.8 by 1.2 inch (20x30.5mm) and is easily installed inside hand-held transceivers. The keyboard is available in the same four styles mentioned above and can be mounted on the side of the transceiver. The Touch-Tone encoder and keyboard, type SME, is priced at $29.95.

In addition to the DTM and SME keyboards and encoders, the major components are also available for amateurs who want to build their own. The keyboard, choice of four styles, is $8.50. The digital Touch-Tone encoder with 1-MHz HC-6/U crystal is $12.50 (encoder with *slim* 1-MHz crystal is $13.50). The miniature printed-circuit board is $2.50. If you purchase a keyboard, encoder and crystal, the PC board, and all resistors and capacitors are provided free of charge.

For more information, write to Data Signal, Inc., 2212 Palmyra Road, Albany, Georgia 31701 or use *check-off* on page 142.

dual-trace oscilloscope adapter

A new RCA dual-tracer adapter that can be attached to any triggered or recurrent-sweep oscilloscope to update it to dual trace operation is now available.

The RCA WM-541A *Dual-Tracer* Adapter provides two displays on a single-trace oscilloscope for simultaneous viewing of two signals. Applications of the new RCA instrument include comparison tests of gain, frequency, response, distortion, phase shift, and time delay. In addition, the WM-541A can also be used to add additional traces to dual-trace oscilloscopes.

Display modes included in the operation of the instrument are channel A only, channel B only, or both A and B channels simultaneously (chopped or alternate). The switching rate is continu-

More Details? CHECK—OFF Page 142

ously variable over a range designed to minimize flicker and beat interference.

The RCA WM-541A has additional features which include ac or dc coupling and vertical position controls for both channels; separate, variable sync-level control with polarity reversing switch; a zener-regulated power supply and LED power-on indicator. The inputs and outputs are terminated with BNC connectors for connection to the oscilloscope The latest cos/mos integrated circuitry is used for high performance operation. The instrument can be used from dc to 10 MHz.

The RCA WM-541A Dual-Tracer Adapter is priced at $108.00. An optional WG-400A Direct/Low Capacitance Probe and Cable is available for $15.00.

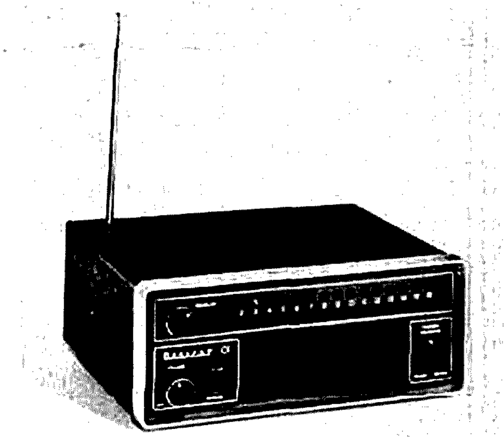
Additional information on RCA Electronic Instruments is available from RCA Distributor and Special Products Division, 2000 Clements Bridge Road, Deptford, New Jersey 08096, or use *check-off* on page 142.

corrosion-resistant vhf antenna

Most mobile antennas include a stainless-steel whip but here is one that is built entirely of stainless steel, brass, and an elastomer compound. This unit has been developed to meet and overcome two significant obstacles to antenna performance – corrosion and the necessity for a ground plane. The construction materials allow this popular model to live happily in a salt environment. The design has no need for a ground plane; this feature allows this unit to operate perfectly on a wood deck or fiberglass trunk lid.

For further details on this high gain, almost indestructible antenna, write to Gam Electronics, Inc., 191 Varney Street, Manchester, N.H. 03102, or use *check-off* on page 142.

synthesized scanning monitor



The new Bearcat 101 is a totally synthesized, five-band scanning monitor featuring a re-programmable custom integrated circuit. In addition to receiving the low (30-50 MHz), high (148-174 MHz) and uhf (450-470 MHz) bands, the unit will also receive the two-meter ham band (146-148 MHz) as well as uhf frequencies from 416 to 450 MHz.

The nerve center of the Bearcat 101 is provided by two exclusive, custom, large-scale ICs: one for *scanning* and the second for a non-volatile *memory* system. With the memory chip, the radio retains all frequencies programmed — without the need for a battery. This feature allows users to order sets fully programmed with frequencies and assures program retention, even if the unit is unplugged or if there is a power outage.

The Bearcat 101 scans 16 channels. Individual lock-out switches are provided for each channel; these are also used in programming frequencies. Channel indicators are light-emitting diodes, providing a scan rate in excess of 20 channels-per-second. Selective Scan Delay, a new feature, permits the listener to remain on a channel for one second longer, in case of a reply on a simplex channel. The Bearcat Selective Scan Delay system permits delay on just those channels desired. Sensitivity in the low and high bands is measured at 0.6 μV;

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on the uhf bands, it typically ranges from 0.6 to 0.9 μV. A six-pole crystal filter offers 70 dB of i-f selectivity.

For more information, write to the Electra Company, Cumberland, Indiana 46229, or use *check-off* on page 142.

precision low-noise op amp

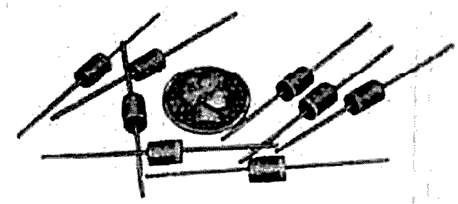
It isn't often that a precision, low noise, ultra-stable, high gain operation amplifier is put into production, and when one is, the cost is usually very high. Not so with a new state-of-the-art amplifier developed by National Semiconductor. Called the LH0044, the new operational amplifier includes all of these features plus low cost.

The LH0044 precision operational amplifier is intended to replace modules and chopper-stabilized monolithic amplifiers and is particularly well-suited for differential mode, inverting, and non-inverting mode applications that require very low initial offset, low offset drift, very high gain and high power supply rejection ratio. In addition, the low initial offset and offset drift of the LH0044 eliminate costly and time-consuming null adjustments.

Specifications include an input offset voltage less than 25 microvolts, long term stability better than ±1 microvolt per month, a maximum offset drift of only 0.5 microvolts/°C, and a noise level lower than 0.7 microvolts peak-to-peak from 0.1 to 10 hertz. Other performance features include a CMRR and PSRR of 120 dB minimum, open-loop gain greater than 120 dB, and a common mode range wider than ±13 volts. The power supply range is from ±2 volts to ±20 volts.

For more information, write to National Semiconductor Corporation, 2900 Semiconductor Drive, Santa Clara, California 95051, or use *check-off* on page 142.

3000-volt silicon rectifiers



Electronic Devices has announced the development of a miniaturized, high voltage, high current silicon rectifier diode with a surge capacity of 300 amperes. The rectifier is an axial lead type. Electrical specifications for the series 3W3 diode are 3000 peak reverse voltage, 2 amp rating with 300 amp surge capacity. Two other similar designs are available with peak reverse voltages of 2000 and 2500 volts. Fast recovery types are also available. The exceptionally high surge capability and small size of these rectifiers results from a special diffusion process and larger junction with lower forward voltage drop.

For complete information, write to the Sales Manager, Electronic Devices, Inc., 21 Gray Oaks Avenue, Yonkers, New York 10710, or use *check-off* on page 142.

tool catalog

A free tool catalog describing over 2500 individual items is offered by Jensen Tools and Alloys. "Tools for Electronic Assembly and Precision Mechanics" is a 112-page handbook of particular interest to amateurs, electronic technicians, engineers, scientists, and instrument mechanics working on fine assemblies. Section headings include screwdrivers, wrenches, pliers, tweezers, files, shears, knives, microtools, relay tools, power tools, metalworking tools, wire strippers, soldering equipment, lighting and optical equipment, work holders, test equipment, engineering and

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drafting supplies and electronic chemicals. New sections include metric tools, books, and wire wrapping tools. A 15-page tool kit section features the world famous Jensen kits for field engineers and kit builders.

Another important feature of the catalog is the inclusion of four pages of technical data on tool selection. These pages include sections on screwdriver selection, machine screw data, tool materials, metal conducitivity, color coding, wire and insulation data, solderability of metals, temperature conversion, drill sizes, metal gauges, metric conversion and safety. Five pages of "Tool Terms" are also included.

A free copy of the Jensen catalog may be obtained by writing to Jensen Tools and Alloys, 4117 North 44th Street, Phoenix, Arizona 85018, or by using *check-off* on page 142.

contact cement

Industrial strength *Zipbond* contact cement bonds most materials almost instantly. It is easy to use, with no pre-mixing necessary, and is used directly from the squeeze applicator bottle (production-line dispenser also available). No heat or pressure treatment is needed, and *Zipbond* sets up quickly at room temperature.

Zipbond is impervious to most chemicals, weather and temperatures. Its bonding strength is not affected by most solvents. Low viscosity allows *Zipbond* to seep into tiny spaces most adhesives can't penetrate. It is very econom-

ical, as one drop covers a 1 inch square area to form a colorless transparent bond. For more information write to Instrument Division, Tescom Corporation, 2600 Niagara Lane North, Minneapolis, Minnesota 55441, or use *check-off* on page 142.

audio power amplifier

A new 40-watt (20 watt rms) B high-fidelity amplifier with total harmonic distortion of 0.2 per cent at 15 watts output is now available from Plainview Electronic Supply. This class B, quasi-complimentary amplifier is capable of delivering full output power into a standard 8-ohm speaker with a 500 mV input signal. Supply voltage can be +36 volts or ±18 volts. Frequency response is from dc to 80 kHz.

The hybrid amplifier is designed for use in communications, stereo, public address and intercom systems, and is priced at $10.65 in small quantities. For more information, write to Bernard Erde, Marketing Manager, Plainview Electronic Supply, 7 Gordon Avenue, Plainview, New York 11803, or use *check-off* on page 142.

transformer catalog

Triad's new *Catalog of Transformers, Inductors, Power Supplies and Circuit Cards*, is now available. The 52-page catalog covers more than 30 categories of transformers, including autoformers, bridging, driver, input, interstage line matching and voltage correction. The inductor section of the catalog lists audio and filter reactors, high Q reactors, tone control and toroidal inductors.

These components are available from 44 Triad-Utrad representatives and distributors worldwide. Catalog requests should be addressed to Steve Fisher, General Manager, Triad-Utrad Distributor Services, 305 N. Briant Street, Huntington, Indiana 46750, or use *check-off* on page 142.

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